ELECTRICAL ENGINEERING

DECEMBER 1946

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1946

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HIGHLIGHTS.........

New Publication Policy. To meet the expressed desires of the membership for a broader and more effective publications service, important changes in AIEE publication policy and procedure will become effective January 1, 1947, in accordance with recommendations made by the publication committee and approved by the board of directors, October 26, 1946 (*pages 576–8*).

Great Lakes District Meeting. In spite of a last minute change in location from Fort Wayne, Ind., to Indianapolis, Ind., as a result of the hotel situation in the former city, the seventh annual meeting of the AIEE Great Lakes District, October 9–11, was carried to a successful conclusion. Featured at the meeting was a general conference on Institute activities and organization of the engineering profession, under the chairmanship of AIEE Vice-President T. G. LeClair of the Great Lakes District, which was addressed by AIEE President J. Elmer Housley (*pages 578–80*). The majority of the papers presented at the meeting were conference or District papers only. Their subject matter included electric machinery, electronics, communications, and basic sciences (*pages 565–9*).

National Electronics Conference. Approximately 2,100 persons attended the second National Electronics Conference which was held in Chicago, Ill., October 3–5. This conference was organized in 1944 as a national forum on electronics developments and their applications to keep abreast of recent rapid growth in that field. The AIEE Chicago Section is one of its sponsors (*page 590*). As an indication of the nature of subjects discussed at the conference, digests of the papers are presented in this issue (*pages 569–74*).

Science and Politics. It is acknowledged generally that the union of science and government during World War II was extremely successful in producing the scientific developments which largely determined the winning of the war. Does the success of this partnership in meeting the abnormal demands of war argue its continuation as insurance of an adequately financed peacetime program? One argument in its favor points out that, inasmuch as the public eventually benefits from scientific research, it has a direct interest in supporting it. This is a debatable subject of immediate and far-reaching importance (*pages 554–6*).

Train Communications. Tests of inductive and radio carrier communications for trains conducted during the past few years under actual operating conditions indicate that such systems, although not entirely satisfactory, are valuable as time-saving devices. The tests were carried out on major United States railroads in co-operation with radio equipment manufacturers, the Association of American Railroads, and the Federal Communications Commission (*pages 547–53*).

Organization of the Engineering Profession. Because there are two aspects to be considered in the organization of the engineering profession—the technical side and the professional side—it would seem that Plan *B* of the four plans submitted by the professional activities subcommittee of the AIEE committee on planning and co-ordination presents the best course to follow. This plan leaves the AIEE and other technical societies free to continue their technical activities and calls for a professional society to carry on all activities of a nontechnical nature (*pages 563–4*).

Slow-Acting Relays. Slow-acting relays have many uses, but their widest application is in automatic telephone systems where they are an essential element. Although little has been said on the subject in technical literature, the advantages of the slow-acting relay, which include small space requirements, low cost, and low energy requirements, far outweigh their only major disadvantage, a question of mounting, and even this has been overcome to a large extent by modern design (*pages 557–63*).

Fault-Current Measurement. A fault-current measuring device must be rapid and positive in action, must not interfere with the electrical characteristics of protective apparatus in the same circuit, and must be prepared to operate automatically. One device which meets these requirements is a magnetic link measuring device which utilizes a surge-crest ammeter and a demagnetizing coil to read the magnetization of the links and to demagnetize them after energization (*Transactions pages 839–43*).

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Radio Relay System. Although designed primarily for military use, an 8-channel microwave relay system which utilizes radio frequencies approaching 5,000 megacycles provides such complete freedom from static and most man-made interference that its basic design principles could be employed profitably in commercial applications (*Transactions pages 798–806*).

Aeronautical Research. The aeronautical research laboratories of the United States are of interest, not only for the part they play in aeronautical developments, but for the variety of electric devices they employ, and the surprising magnitude of their electric power loads. For example, the Ames Aeronautical Laboratory at Moffett Field, Calif., one of three laboratories of the National Advisory Committee for Aeronautics, will have more than 150,000 connected horsepower when wind tunnels now under construction have been completed (*Transactions pages 833–9*).

Teaching Electricity. To comprehend fully the significance and limitations of each of the relations, formulas, and rules which make up the body of a science, it is necessary to understand the basis upon which each was obtained. Thus, a presentation of most of the fundamental relations in electricity and magnetism in general form, with the basis of each relation indicated clearly, should provide the beginning student with the means for acquiring a fundamental understanding of electricity and magnetism, and forms his basic training in engineering analysis (*Transactions pages 828–33*).

Corrections. Attention is called to the following misprints occurring in recent issues of *Electrical Engineering*. In the paper "The Influence of the Concentration and Mobility of Ions on Dielectric Loss of Insulating Oils" by Bun Po Kang, published in the Transactions section of the July 1946 issue, on page 403, column one, fourth line from the bottom, "within a limited rage of temperature" should be corrected to read "with a limited range of temperature," and equation 3 on page 404 which appears as "$W=A-B/T$" should read "Log $W=A-B/T$." On page 458 of the same issue, also in the Transactions section, an error appears in the paper "Operating Experience With Distance Ground Relays" by W. A. Wolfe. The second term of equation 2 should read $I_{0GF}\dfrac{Z_{0GH}-Z_{1GH}}{Z_{1GH}}$ instead of $I_{0GF}\dfrac{Z_{0GH}+Z_{1GH}}{Z_{1GH}}$ as printed. In the paper "Application Ratings of Indoor Power Circuit Breakers" by O. B. Vikoren, published in the November issue, the 14th line, second column, page 270 of the Transactions section, which reads "in the 400-ampere and 440-ampere tests" should be corrected to "in the 400-ampere and 550-ampere tests."

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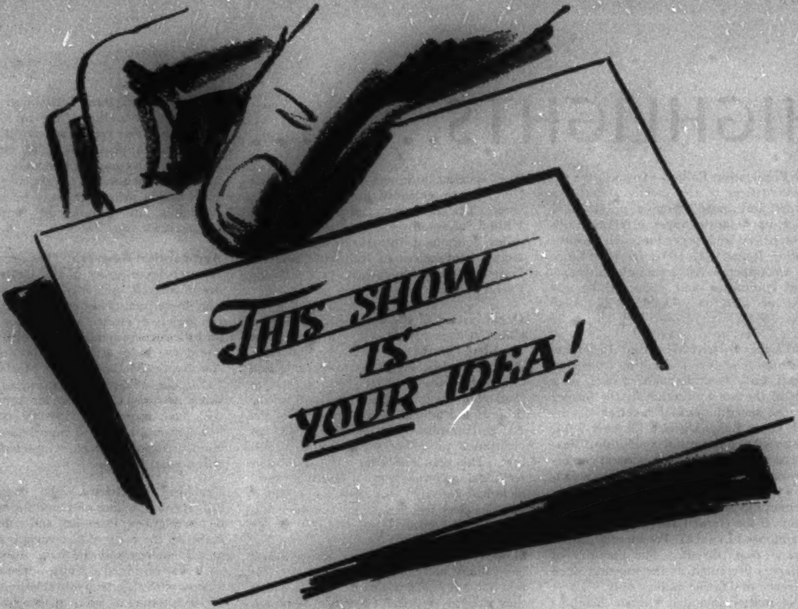

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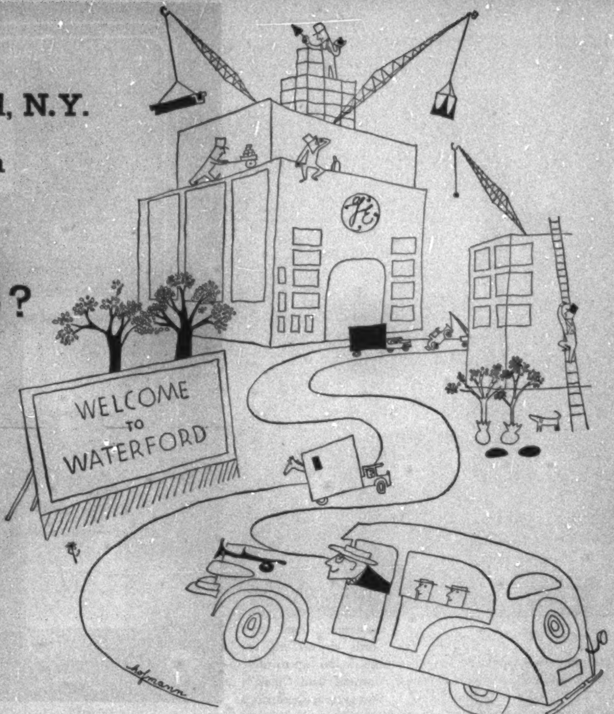

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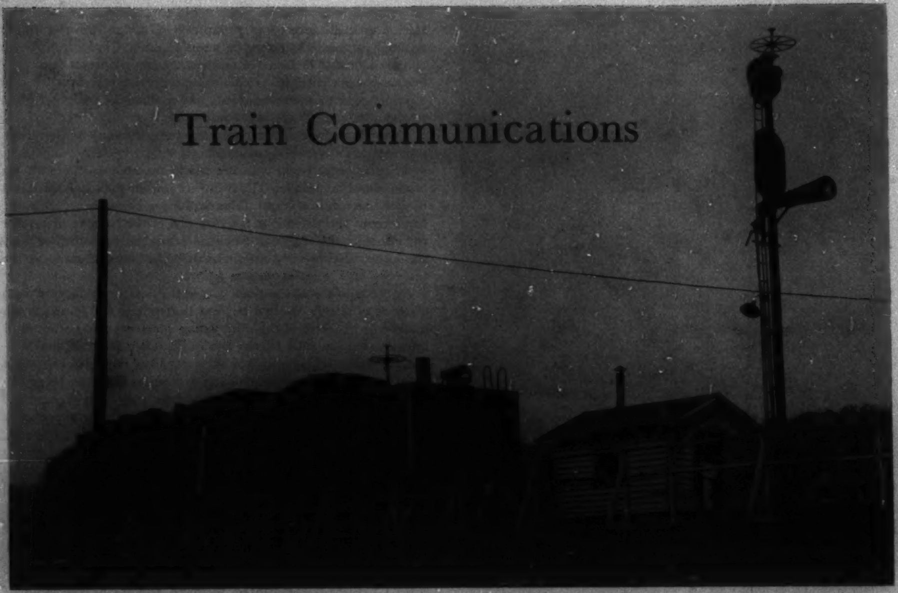
Train Communications



During the past few years many railroad companies have conducted extensive tests of inductive and radio carrier communication systems. Excerpts from reports of these tests point out that these systems though not perfect are already rendering useful service.

ALTHOUGH radio communication for use by railroads was conceived several years ago,[1] it has not come into extensive use for several reasons, among which is the hesitancy to adopt any system until all adverse operating characteristics are eliminated. The Association of American Railroads, the radio equipment manufacturers, and the Federal Communications Commission, with the co-operation of the major railroads of the United States, have conducted during the last few years very extensive tests of both inductive and radio carrier communications. These tests have been conducted under actual operating conditions by installations on regular established scheduled trains, and have formed the best possible basis for selection of a suitable communication system.

The following is a symposium from the reports on these tests.

The Atchison, Topeka and Sante Fe Railroad in June 1944 conducted tests of end-to-end communication on a 3,500-ton freight train on a run from Bakersfield, Calif., to Chicago, Ill. For these tests a modified turnstile type of antenna was mounted on the top of Diesel locomotives and in the case of steam locomotives, on the top of the tender. An antenna of the same type was mounted on the roof of a business car attached to the rear of the train. The reception at both ends was unaffected in traveling over all types of terrains, in mountains, deep cuts, grooves, and under steel over-structures. Varying atmospheric conditions, including two rain and electric storms, were encountered during the 2,200-mile trip, but had no effect on the communication. The maximum distance of transmission of about two miles was accomplished at times when the engine was uncoupled from the train and moved to a roundhouse or to service tracks. Quality and volume of transmission was satisfactory at all times except for background noise from the equipment located on the steam engine which, while not desirable, was not sufficiently objectionable to interfere with transmission.

The Denver and Rio Grande Western Railroad has conducted extensive tests of end-to-end and point-to-train communication in which radio was used on frequencies ranging from 30 to 2,700 megacycles. The following is part of the report to the Federal Communications Commission, dated December 8, 1944.

Despite the adverse conditions of terrain and extensive tunnels, the results of the tests to date are exceptionally encouraging. Early in the tests it became evident that the greatest source of

---

Material for this article was obtained from the Association of American Railroads, New York, N. Y.

Figure 1. Inductive carrier station aboard a Seaboard Railway engine

communication difficulties would be in the tunnel area, a 41 mile stretch, where the Rio Grande operates over the Denver & Salt Lake Railway line, starting 15 miles west of Denver and extending through the Moffat Tunnel. In this territory there are 30 tunnels varying from a few hundred feet in length to the Moffat Tunnel which is 6.2 miles long.

In this territory due to the frequency of the tunnels there are numerous times that either the locomotive or caboose, or both, are in a tunnel at the same time. It is in this extremely rugged territory with difficult operating conditions that the use of radio as a reliable means of communication would introduce an added factor of safety and materially aid in the handling of trains and expediting the movement of traffic.

In conducting caboose-to-locomotive communication tests, there were attempts to use different types of equipment varying in receiver sensitivity, transmitter power, frequency, and method of modulation. All equipment is installed in freight service—one unit on a Diesel locomotive, the other on a caboose—and tests are made between front and rear of the freight train on the main line between Denver and Salt Lake City or some intermediate point.

The Northern Pacific Railway conducted tests on radio equipment during August 1945.

Although it was fully anticipated that some blind spots would be encountered due to curvature and foliage especially in mountain areas, the only dead spots we actually encountered were tunnels. Good communication was obtained everywhere except when one or both ends of the train were in a tunnel and in every case when the train was long enough or the tunnel short enough, communication was established around the tunnel as soon as the engine emerged and the caboose had not yet entered.

On October 17, 1945, end-to-end and point-to-train communication tests using both inductive carrier and radio equipment were conducted on the Illinois Central Railroad between Freeport, Ill., and Waterloo, Iowa.

Radio communication between the ends of the train could be effected at all times. As the train approached the radio station at Jesup, Iowa, attempts were made to communicate with this station at a distance of approximately 40 miles from the station without success. At points between 19 and 11 miles from Jesup the station could be heard intermittently, but consistent communication was not possible. Good communication was established at a point about 11 miles from Jesup and maintained from that point to about the same distance beyond Jesup.

As the train was approaching the Jesup station, arrangements were made for a special test as follows:

The station at Jesup was directed to transmit continuously for a considerable period. The radio station on the locomotive was directed to transmit also. Observations were made on the radio receiver on the test car at the rear end of the train to determine the effect when the signals from the Jesup station and those from the locomotive reached approximately the same intensity at the test receiver. Results were as follows:

As the rear end of the train passed the Jesup station, and for a distance of a mile or two beyond that point, the Jesup station maintained control of the observation receiver, and no signal or interference was heard from the radio transmitter on the locomotive which was about 3/4 mile ahead of the test car. At a point somewhere between one and two miles from the Jesup station the transmitter at the front end of the train took over control of the receiver. This was accomplished without any appreciable flutter or presence of both signals. The train, during this test, was moving at a fairly rapid speed. Although this test was very limited in scope, it seems probable that, in the case of a train moving at normal speeds, the area in which communication would be obscured by flutter because of approximately equal signal intensities received from two wayside stations will probably not be great.

During April 1945, the Texas and New Orleans Railroad Company conducted tests on the inductive type of communication, on end-to-end, point-to-train, and train-to-train very-high-frequency radio communication.

The first train started from El Paso, and made a run with tonnage freight to Valentine (161.5 miles east). On this trial run it was expected that satisfactory communication would be obtained at all times between the locomotive and caboose and with a fixed station whenever the train was not more than ten miles distant. The distance that the train did communicate with the station varied and under favorable conditions contact was made up to twenty miles between the train and one of the fixed stations. For this test the first train left Valentine for a return run to El Paso and the second train left El Paso for Valentine, it being the intention to test the radio equipment between two trains approaching, meeting, and leaving each other. The first satisfactory contact between the two trains was established when the trains were about six miles apart and was maintained continuously until the trains met and were again about five or six miles apart. The terrain between El Paso and Valentine is generally mountainous, the railroad usually traversing the low portion between high hills and mountains which affords a very severe test of radio equipment of the frequency used because of the general line of sight characteristics at this frequency.

The Seaboard Air Lines Railway conducted a test run in 1944 between Richmond, Va., and Savannah, Ga., using inductive equipment.

Transmission was loud, clear, and of good quality. When passing a 110-kv power line no interference was noticed.

No difficulty was encountered in going around curves, through cuts, and so forth. At one time the train passed through a steel truss bridge with steel members on the bottom side and top. This did not seem to affect transmission in the least.

An inductive radio test was made on the Southern Pacific in 1944 from moving train to fixed stations and from head to rear end, in which a 73-car freight train was operated over a distance of 180 miles. This train originated at Roseville, Calif., and ran through to Fresno. When running in parallel to the telegraph line carrying the Western Division wires we were able to communicate from moving train to Stockton (42 miles). The volume and quality improved as the train proceeded to Stockton; however, it was audible and understandable at all times. The country through which we traveled was flat and level and telegraph line varied in distance from center line of the track from 50 to 300 feet, but the greater portion of the line averaged about 50 feet from center line of the track. The maximum distance that we talked from moving train to fixed station was some 70 odd miles. The test was quite satisfactory in every respect. We experienced some slight interference where wig-wags were operating in yards, but these noise levels did not interfere with the voice to the extent that it could not be clearly understood.

The Chicago, Milwaukee, Saint Paul, and Pacific Railroad, during December 1944, conducted tests of end-to-end and point-to-train communication using the inductive system.

Diesel locomotive and a mail and express car were equipped with mobile sets. At the Beloit Division office a similar set was installed.

On the westbound trip communication was carried on with Beloit satisfactorily from about 50 to 60 miles east to about 20 to 30 miles west of that point. Eastbound the range was from 30 to 40 miles west to 60 miles east. Electrical interference attributed to the auxiliary and main generators of the locomotive interfered seriously with the reception on the locomotive at certain engine speeds and loads. The effect of this interference was to render two-way communication practically impossible. The application of filters to certain parts of the locomotive electrical equipment is planned.

The New York Central Railroad during April 1945, conducted tests of end-to-end and point-to-train inductive carrier communication.

These installations operated on 175 kc, employing frequency modulation. Loudspeakers were provided for calling and handsets for talking and listening although both the handset receiver and loudspeaker could be used simultaneously by depressing the handset hook.

Transmission between all fixed stations was of good quality, quiet, and of ample volume. Communication between the mobile units was also good except that when wayside wires were below track level communication, particularly to and from the engine, was noisy. On one test when the engine was uncoupled from the train communication between engine and caboose was carried on over a distance of six miles. Communication between the mobile units and fixed stations was also satisfactory; the maximum distance over which such communication was attempted was 27 miles.

INDUCTIVE COMMUNICATION

Inductive train communication consists of communicating between mobile units (locomotives and cabooses) and between mobile units and fixed stations by means of carrier currents which are superimposed on the usual telegraph and telephone wires which parallel the right of way. Communication is by telephone, two-way, usually simplex operation, and as far as the user is concerned is similar to radio. One or more than one channel can be provided. It in no way interferes with the normal use of these telegraph and telephone wires.

For mobile units coupling is usually obtained by means of vertical plane loop antennas in line with the track. On transmitting, these loops are energized with the carrier frequency modulated by the voice signals, and induce, by electromagnetic induction, such currents longitudinally into all wires on the pole line. Thus these induced currents flow in all wires and appear as a voltage between all wires and ground. On receiving the carrier currents flowing in the line are induced into the loop. Some systems use the same loop for both transmitting and receiving, others use a separate smaller loop for receiving, and if there are two channels, use a separate receiving loop for each frequency.

At the fixed station the set is connected between ground and one or more wires on the pole line through a coupling unit, which consists essentially of small capacity condensers. While coupling is usually made to one wire, or to the two wires of a telephone circuit only to keep the pair balanced for telephone service, the capacity between wires on the pole line for these high frequencies, makes this essentially coupling to all wires on the pole line. Tests have indicated there is no advantage gained coupling the way station to more than one or two wires.

On some tests the coupling on mobile units has been made by means of a capacity-type antenna, and coupling to the line wires obtained by electric induction. However, for transmitting the loop antenna has become almost universal, while the capacitance-type antenna is used in some cases for receiving.

Certain of the earlier systems used the upper side band of a 5,700-cycle carrier. Amplitude modulation at 75 and 80 kc has been used and frequency modulation from 70 to 200 kc. The usual frequency modulation deviation ratio used is one-to-one, which occupies a band about 6 kc wide 3 kc above and below the carrier frequency. Manufacturers state that they can supply equipment to operate from 50 kc to 250 kc, and apply frequency modulation.

Frequency modulation, and the use of the higher carrier frequencies, has the advantage of reducing extraneous noise.

By far the greatest loss is the coupling loss between the mobile units and the pole line. It will increase with the separation of the pole line from the track, and if the pole line is above or below the level of the antenna on the mobile units. The coupling loss is less the higher the frequency. Data are very meager and because of the large number of variables would be difficult to set up. Data obtained from one manufacturer, at 170 kc, show the coupling loss between one mobile unit and the pole line to be between 25 and 45 decibels for separations from 30 to 200 feet.

The coupling loss between the fixed station and the line is practically zero.

The loss along the pole line will increase with the frequency; representative values given were 0.4 decibel per mile at 70 kc and 0.8 decibel per mile at 170 kc, although on some lines it has measured as high as 1.2 decibels per mile. It will be increased to a certain extent by the presence of iron wire and bridged taps, particularly if these bridged taps are long and go into lead-sheathed or underground cable. The effect of bridged taps can be eliminated by the use of blocking filters and in some cases this may be necessary.

The Kansas City Southern Railway demonstrated the use of carrier telephone equipment on June 21, 1944.

> The system is a carrier system and 80 and 175 kc were used in operation. The system extends from Kansas City, Mo., to Shreveport, La. (560 miles). At present one engine and two cabooses are equipped. Shortly it is expected five engines and five cabooses will be equipped. There are 17 fixed stations along the line from two to 62 miles apart. The system is used for station to station, station to train, and front to rear end communication. The carrier is superimposed upon a telephone pair which, in addition, provides a voice channel, one simplex telegraph channel, and a 22-kc telephone and telegraph carrier system, the latter providing one long distance telephone and two 2-way telegraph channels.
>
> The set on the engine is connected to a loop consisting of seven turns of number 12 copper wire. During the test the loop used was mounted on top of the tender in a vertical plane and was approximately six feet high and 12 feet long. At other times a loop arrangement had been tried on the engine consisting of one turn on each side of the boiler. The top wire was supported about one foot away from the boiler and about on a level with the sand dome. The bottom wire is just below the runway. These two loops, one on each side of the engine, were connected in series to give the effect of a 2-turn loop.
>
> So far as reception and transmission are concerned, either arrangement appeared to work satisfactorily.
>
> The loop on the caboose consisted of four turns. All four turns were carried across the top of the caboose in the center. On the bottom two of the turns were carried on each side and just below the framework of the caboose.
>
> The transmitter output of 75 watts at carrier frequency is fed into the loop of either the engine or the caboose.
>
> The system operates from the fixed station by inducing a magnetic field around the paralleling wire line which cuts the turns of the loops on the moving equipment and induces voltages therein. These voltages feed into the receivers on the moving equipment.
>
> On the moving equipment the transmitters circulate currents through the loop and these currents generate magnetic fields which in turn cut the wire line and induce voltages therein. At the fixed station these voltages act directly on the receiver. They also produce current in the paralleling line wire and this current generates a second field which interacts with the loops on other units of mobile equipment.
>
> The arrangement between fixed station and mobile equipment is somewhat analogous to a 2-winding transformer. The arrangement between mobile units is somewhat analogous to a 3-winding transformer, winding number 1 corresponding, say, to a caboose, winding number 2 to the paralleling wire line and winding number 3 to the equipment on the engines.
>
> One company has advised that they have found that when the loops are oriented in the vertical plane they provide the most efficient transfer of energy to the wire line when the latter is centered and perpendicular to the loop. For a satisfactory transmission the wire line should be not more than 125 feet away from the track.

RADIO COMMUNICATIONS

The technical methods and equipment used varied from company to company. The Denver and Rio Grande Western Railroad reports

> To date four different types of radio equipment have been tested in actual service tests:
>
> 1. Frequency-modulation 60-watt emergency equipment operated in the 35-megacycle band.
>
> 2. Phase-modulation 15-watt experimental unit of the Navy in the 2,600-megacycle band.
>
> 3. Frequency-modulation 25-watt repeater equipment on a frequency of 117.65 megacycles.
>
> 4. Air-bourne amplitude modulation 6 watt *SCR-624A* unit on a frequency of 156.525 megacycles.
>
> A number of antenna arrangements have been attempted. The vertical quarter wave to ground, vertical folded dipole, horizontally polarized dipole, *J* type with matching stub and compensating capacitance, and an array composed of a driven vertical quarter wave with a parasitic reflector and director.
>
> The 60-watt frequency modulation emergency equipment proved very satisfactory everywhere except in the Moffat Tunnel. The test was conducted on a freight train, caboose-to-locomotive two-way from Denver to Salt Lake City and return. We did not encounter any noise or interference from man-made static. No fading, flutter or distortion was noticed throughout the trip. The 15-watt phase-modulation experimental unit was a one-way, locomotive-to-caboose test, mounted on a freight train, and tested from Denver to Salt Lake City. It was possible to hold conversation in one direction, but as the system of testing had to be interrupted, it was not deemed advisable. Field strength was measured at the receiver and a continuous wave audio note was modulated on the carrier for the benefit of a loud speaker in the caboose. Because of a radio equipment failure, and the brief time the equipment was available to us, no test was obtained in the "shelf" area where the track runs for some distance on the side of a mountain with no mountains or other possible reflecting surfaces on the other side of the track. It is possible that this area could give trouble at this frequency. No flutter or distortion was apparent. Fading, of course, was registered on the meter, and in a few cases went below the level where the automatic voltage control could maintain normal speaker volume. In one case, with the locomotive in a tunnel, the signal died because of lack of reflecting surface. It is thought that a parasitic reflector could be used to re-radiate the wave. An interesting fact was that radiation was more effective in the six mile long Moffat Tunnel than out of it, the tunnel apparently acting as a wave guide. No flutter or distortion was apparent throughout the test.
>
> The 25-watt frequency-modulation repeater equipment did not perform in any of the tunnels, and in some cases we had trouble obtaining enough volume in cuts, canyons, and so forth, to properly operate the speakers. This trouble was no doubt due to the fact that this equipment uses such a low-grain receiver. No flutter or distortion was evident except when we used a horizontally polarized dipole antenna. No noise or interference from man-made static was encountered. All tests with this equipment were conducted on a freight train, engine-to-caboose, two-way between Denver and Grand Junction, Colorado.
>
> The air-bourne units are now being tested in freight service, engine-to-caboose two-way between Denver and Grand Junction. No reception has been obtained in the tunnels, consequently giving us very poor service in the tunnel area. The only time interference is noticed is when the units are in close proximity to motor cars, probably due to the magnetos used for ignition. Under this condition some flutter becomes evident but the signal is still intelligible.
>
> In comparing the signals received by the different units, the re-

ceivers used in each case must be considered. The sensitivity of all the receivers used is of the order of three microvolts for a ten decibel signal-to-noise ratio, except in the case of the 25-watt repeater units which have a rather low gain.

In the case of the antenna experiments, the best results to date have been obtained from the use of the J type and the quarter-wave vertical to ground. We do not consider the horizontal type of antenna satisfactory for this type of radio communication. The other types of antennas used varied somewhat in the signal strength received, but no harmful effects were noticed.

Table I. Railroad Radio Authorization by Federal Communications Commission (to July 15, 1946)

Railroad	Transmitter Location	Nature of Installation
The Atchison, Topeka and Santa Fe	Chicago, Ill.	Yard—fixed station
	San Francisco, Calif.	Yard—fixed station
	(Portable and portable-mobile)	14 stations
	Chicago, Ill.	Fixed station
The Baltimore and Ohio	Newcastle, Pa.	Yard—3 units
Chicago, Burlington and Quincy	Chicago, Ill.	Yard—fixed station
	(Mobile)	Train stations—30 units
Chicago, Milwaukee, Saint Paul and Pacific	(Mobile)	Train stations—15 units
Chicago, Rock Island and Pacific	(Portable and portable-mobile)	Experimental
	(Portable)	Train stations—7 units
	(Portable)	Yard and Terminal
	(Mobile)	14 units
	(Mobile)	Train stations—20 units
The Denver and Rio Grande Western	(Portable and portable-mobile)	Experimental
	(Mobile)	Train stations—32 units
	Denver, Colo.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—7 units
Florida East Coast Railway Company	Buena Vista Railway Yards, Miami, Fla.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—11 units
Gulf, Mobile & Ohio	Meridian, Miss.	Yard & terminal—fixed staton
Jacksonville Terminal Company	Lee St. Tower, Jacksonville, Fla.	Terminal—fixed station
	Myrtle Ave. Tower, Jacksonville, Fla.	Terminal—fixed station
	(Portable-mobile in vicinity of Jacksonville)	Terminal stations—12 units
Missouri Pacific Railroad	(Mobile)	Train stations—32 units
New York Central	Cheektowaga Township, N. Y.	Yard—fixed station
	Bethlehem Township, N. Y.	Yard—fixed station
	Manlius Township, N. Y.	Yard—fixed station
	(Portable and portable-mobile)	Yard—24 stations
	(Portable and portable-mobile)	Yard—12 stations
	Hammond, Ind.	Yard—fixed station
	(Mobile)	Yard—5 units
Northern Pacific Railway Company	(Mobile)	Train stations—5 units
Pere Marquette Railway Company	Grand Rapids, Mich.	Train Station—fixed
	(Mobile)	Train stations—5 units
Seaboard Air Line Railroad Company	Hamlet, N. C.	Yard and terminal—fixed station
	Hamlet, N. C.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—fixed station, 15 mobile units
Union Pacific Railroad Company	Fairfax Railroad Yard, Kansas City, Kans.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—16 units

Train radio tests were conducted on the New York, New Haven, and Hartford Railroad in the period September 12 to December 15, 1944, utilizing frequency modulated equipment operating at a frequency of 30.66 megacycles. Three sets were constructed. In its final form, each set of equipment consisted of a transmitter, a power supply, and a receiver, shock mounted in a dust-tight weatherproof steel housing. A handset microphone of the push-to-talk type was used in each transmitting station. Voice reception was by either the handset or a loud speaker, the loud speaker serving to announce the initiation of a call.

The two mobile stations had an output of 30 watts each and the fixed station 60 watts.

The antennas on the mobile equipment were in the form of a U-shaped pipe railing mounted on the tender just behind the coal pile and on the caboose roof alongside the cupola. In all cases the antenna was entirely within the clearance limitations of the equipment and was effectively grounded to provide protection from possible accidental contact with overhead trolley wires in the electrified territory. For the fixed station, an antenna in the form of a large ring located on top of a grounded steel pole was used.

It was found that the range of the radio equipment was not fixed. It varied with the height of the antenna, condition of terrain, and other factors which caused shielding of the radio signal. However, based on tests at several locations, it was found that satisfactory operation could be obtained between the fixed station and mobile units up to distances of approximately 10 miles and between mobile units at separations up to approximately 5 miles.

RELAY SYSTEMS

It is likely that radio relay systems will find application in the railroad communication field where there is a sufficient volume of traffic to justify the cost of installation or where unusual conditions cause frequent interruptions to continuity of service rendered by wire lines and the cost of maintenance of such lines is unusually high. It should be pointed out, however, that it now appears the cost of radio relay systems will be so high that they probably cannot be economically justified except in situations where they can be operated with fairly heavy loads.

In the railroad field it appears that there may be a possibility of operating relay systems in conjunction with train-to-point radio systems where it will be possible to utilize common equipment for power supply, antenna supporting structures, auxiliary power units and housing facilities, which will tend to make for more economic installations of both types of equipment and may make possible the utilization of radio relay links under conditions where traffic loads would not ordinarily justify such installations.

So far as is known, the only railroad installation to date of point-to-point communication by use of ultra-high frequency is an experimental link set up between Kansas City, Mo., and Topeka, Kans., by the Rock Island Lines.

EQUIPMENT

As a result of the subject tests equipment designed for railroad use in the 158–162-megacycle band has been developed by several manufacturers. Equipment used in the earlier tests was police, aviation, and military radio equipment adapted for test use in connection with railroad operations. As might be expected some failures of this equipment occurred. Attempts have been made in the new equipment to eliminate the weaknesses exposed and develop rugged equipment capable of dependable operation in railroad service. One requirement made clearly evident by the tests was the need for sensitive receivers in order that satisfactory communication may be obtained where mountains, deep cuts, or foliage, may cause severe attentuation of received signals.

Table II. Induction Installations

Date in Service	Railroad	Type Equipment
Mar. 1937	Bessemer and Lake Erie	2-way
Mar. 1940		
July 1940	Cleveland, Cincinnati Chicago and Saint Louis	1-way
May 1943	Chicago, Burlington and Quincy	1-way, 2-freq.
Mar. 1944	Chicago, Burlington and Quincy	1-way
May 1943	Great Northern	1-way
Oct. 1940	Louisville and Nashville	1-way
Apr. 1942	Norfolk and Western	2-way
Mar. 1943	Pennsylvania Railroad	1-way, 2-freq.
Dec. 1942	Pennsylvania Railroad	2-way
Nov. 1940	Pennsylvania Railroad	1-way
May 1942		
June 1943	Pennsylvania Railroad	1-way, 2-freq.
Mar. 1942	Pennsylvania Railroad	1-way
July 1943	Pennsylvania Railroad	1-way
Aug. 1944	Pennsylvania Railroad	2-way
Feb. 1945	Terminal Railroad Association of Saint Louis	1-way
	Kansas City Southern Railroad	2-way
(Proposed)	Atlantic Coast Line	2-way, 2-freq.
(Proposed)	Chesapeke and Ohio	2-way
(Proposed)	Pennsylvania Railroad	2-way, 2-freq.

Adequate automatic control of receiver output volume is another important requirement.

The latest transmitting equipment developed for this service generally has a power output of from 5 to 50 watts, radio frequency range of 152 to 162 megacycles, audio frequency range of 300 to 3,000 cycles, and is usually designed for use with 52- or 72-ohm loads.

The inductive carrier equipment, having been under development for railroad use over a longer period, has been standardized to a greater degree. Power output of this equipment is uniformly 5 watts for wayside stations and 50 watts for mobile stations, employs carrier frequencies between 88 and 175 kc, and frequency modulation with a 1-to-1 deviation ratio.

Radio equipment used in the various tests employed both amplitude and frequency modulation. It was found during these tests that noise from electrical sources caused very little trouble in reception of signals above 150 megacycles. While it was apparent that frequency modulation resulted in greater exclusion of interference from other stations on the same frequency, and with amplitude modulation heterodyne interference occurred, there has been insufficient operating experience in regular service to determine whether this discrimination characteristic of frequency modulation would be an advantage or disadvantage.

Little difficulty has been experienced in teaching railroad operating personnel to use either radio or inductive carrier equipments. In general the equipment has a minimum of external controls. Operation of a "push to talk" button on the microphone is the only control used in normal operation of the equipment.

The main problem in connection with antennas for both radio and inductive carrier has been to develop efficient antennas for mobile use within the limitation as to height imposed by the railroad clearance pattern. On frequencies above 150 megacycles horizontal antennas caused variation or "flutter" in the received signals. Consequently most radio antennas for mobile service have been vertical quarter-wave or variations thereof. The most used variations are top loaded vertical antennas, usually with a ground plane of the cartwheel type, and modifications of the coaxial type of antenna.

Fixed station antennas, although uniformly vertical, have varied rather widely in other respects. Some directional arrays have been used in attempts to extend the range of point-to-train radio communication. Considerable development work must be done in connection with fixed station radio antennas.

In inductive carrier communication, loop antennas, oriented in a vertical plane parallel to the tracks, have been used almost entirely. The particular form of loop varies, sometimes being entirely insulated from the mobile unit on which it is installed and sometimes being connected to the unit in such a manner that the output is fed inductively into the rails.

In end-to-end communication adequate volume and quality of signals have been maintained in the various tests with the exception of a few isolated instances where some flutter occurred in territory including deep cuts and foliage. The tests have indicated that radio communication in the 158–162 megacycle range fails in tunnels although some efforts have been made toward development of methods of overcoming this difficulty. One test involving tone-modulated 2,660-megacycle equipment was successfully received through the 6-mile Moffat tunnel but failed in a shorter tunnel because of lack of reflecting surfaces.

Inductive carrier communication in end-to-end service has been equally successful except where wayside wires are too far from the track or too far above or below track level. The maximum permissible distance between wayside wires and track averaged 200 feet. Greater attenuation was noted when wires were below track level than when considerably above track level.

Good communication over distances of approximately 5 miles by radio and approximately 15 miles by inductive carrier was obtained in end-to-end and train-to-train service. In point-to-train service the average distance was 12 miles for radio and 50 miles for inductive carrier. Results of the tests show that line-of-sight conditions are generally necessary for reliable radio communication in the 158–162 megacycle band, although in some cases communication ranges well beyond line-of-sight are realized.

USES

End-to-end communication saves substantial amounts of time on many operations. Radio communication is not reserved for emergencies, but is used extensively in normal operations. For example

1. The engineer may be notified when proper train line pressure has been reached at the rear of the train.

2. The engineer may be notified as soon as the rear of the train clears crossovers and switches, thus permitting full speed sooner than would otherwise be possible.

3. Increased efficiency may be obtained in detecting and remedying hot boxes.

4. The engineer may be notified of the flagman's return to the train, thus saving time and eliminating uncertainty.

5. Control of Diesel-electric locomotives may be facilitated through more effective co-ordination in manipulating controls.

6. Unusual conditions may be discussed by the conductor and engineer.

7. The engineer may be advised of the time to start and stop at specific points.

8. The engineer may be advised of sticking brakes.

9. The conductor may advise the engineer to stop the train, thus avoiding hazards incident to application of air from rear of train. (Application of the airbrakes from the rear of the train, which might be necessary were it not for radio communication, may part the train and cause delay to other trains.)

10. The engineer may advise the conductor where water is to be taken, when a train to be met is in clear on the siding, or when difficulties develop with the engine.

11. The conductor may advise the engineer when the rear of the train is nearing switch so that the train may be stopped close to switch to pick up the trainman after lining the switch.

12. The conductor may advise the engineer when the brakes are released on the rear of the train.

13. The conductor may advise the engineer when to stop the train to spot cars to be loaded, unloaded, or re-iced.

14. The conductor is often able to notify the engineer when fire is dropping from firepan.

15. The engineer may notify the conductor of the indication of train order and interlocking signals.

16. If the train should break in two while in a siding the conductor may instruct the engineer as to the proper handling to couple.

17. If the conductor should miss a message handed up at a station, the engineer may read him the contents from his copy, or vice-versa.

Numerous advantages are also available from point-to-train communications.

1. The dispatcher may advise the train on what track cars are to be set out before the train arrives.

2. The conductor may advise the dispatcher when a train he is meeting leaves a station.

3. The dispatcher may inform the conductor of unscheduled train and designate the nearest siding.

4. The dispatcher may tell the crew what work is to be done when they arrive at station.

5. The conductor may advise the dispatcher of low steam and delay in schedule. Necessary orders may then be issued and repeated.

6. The dispatcher may advise the train where to wait to avoid being delayed by an opposing train.

7. The conductor may advise the dispatcher that the train has broken in two and is delayed.

8. The conductor may advise the dispatcher when the train has departed.

The Denver Rio Grande Western Railroad reports that radio communications aided them on their test run.

In one case, the radio train approached Grand Junction from the east and learned that the yards were congested. There are two ways of entering Grand Junction—one on the main line, and one from the south with switches from the main line about a mile and a half east of the main line entrance. The latter route leads to the east (or storage) yard, which is used for train handling only when the main yard is congested. The usual method of handling this situation is for the head end brakeman to walk back to the caboose and inform the conductor of the situation, and then return to the middle of the train to relay the hand signals from the rear to the front. The train is then backed up $1^{1}/_{2}$ miles into centralized traffic control territory to clear the switch which will take the train into the yards on the alternate route. This takes considerable time because of the walking involved and the caution with which the train must be moved back. In the case of the radio equipped train, the engineer informed the conductor immediately of the condition in the yards, the conductor ordered the train to back up, and he called signals over the radio. It is estimated that the radio saved about an hour and 15 minutes in handling the situation, besides avoiding the possible detention of any other train that might have been close behind.

On another westbound trip the centralized traffic control headed the train into the passing track at DeBeque, Colo. There was a broken rail between DeBeque and Akin and the dispatcher allowed trains to proceed on a permissive card only. We filled the passing track at DeBeque. There were four passenger trains behind us, and they could not proceed to Akin without "sawing" by the congested passing track. The dispatcher asked us to break our train and put the front half on a stock track adjoining the passing track so he could send the extra train at Akin to occupy our track with us. This would allow a clear track at Akin for the four passenger trains to go through. This was a complicated move and as our orders came from the telephone near the head end, the information had to be relayed back to the conductor on the caboose. His acknowledgment was received and instructions as to how to handle the cut given by him. The move was made without the conductor having to leave the caboose. No walking was involved except for the head brakeman to walk back 20 cars to effect the cut. The trainmen suggested that in this case probably an hour's time was saved by the use of the radio.

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Science, Politics, and the National Welfare

FREDERICK L. HOVDE

Originally presented as an address before the recent National Electronics Conference in Chicago, this article considers one viewpoint in the current controversy over the question of government support of scientific research. The author takes his stand on the side of federal promotion of science, provided the policy to be pursued is both scientifically and politically acceptable. Undoubtedly, some readers will take issue with him. Truly, here is food for serious thought, and potential for extended debate.

MY AUDIENCE TODAY is composed of scientists, engineers, educators, and executives, all of whom are concerned professionally with that special and highly useful branch of science called "electronics." You are met to discuss the fate of this science, to speculate on its future, to discuss its problems, to exchange information, and to make plans enabling your science to increase its service to society.

Such conference objectives are laudable and worthy of your time and attention as active scientists. But you are also well-educated and responsible members of a society whose welfare depends primarily on how wisely it is governed and how well it uses the fruits of your work. Here is a problem far more important and far more difficult than any technical problems you will discuss in the conference sessions which follow.

This problem must, I believe, receive your constant and serious attention; otherwise your scientific labors will be wasted—even worse, will be used to destroy a civilization potentially finer than any in the past because man at long last has the power to control the forces of nature. Highly doubtful, but more important, is the question of whether or not man has the power and will to control himself.

UNION OF SCIENCE AND POLITICS

Sigmund Freud, the great Viennese scientist, founder of psychoanalysis and contributor to modern psychiatry and psychology, once remarked that man had three impossible tasks—namely, to educate, to govern, and to heal. This pessimistic observation is uncomfortably near the truth, as a little reflection on the history of man's progress with these tasks will show.

We cannot, however, accept Freud's dictum. To accept defeat before the battle is to be defeated for sure. We must continue relentlessly with all the genius and hard work at our command to attempt the solution of the problems of educating, governing, and healing. Even if Freud was right, we must continue our search for the answers. The intellectual and spiritual activity involved is, by its very nature, justification enough.

In the title of this article I have coupled the words "science" and "politics." These bedfellows are strangers. Some of you may think that they are incompatible, and, therefore, not likely to produce much. Whatever your thoughts about this union, they are joined together for better or for worse in both our national and our international life. The scientist is perhaps more surprised than the politician, who long has been used to strange bedfellows. In preparing these remarks I thought not only of the commonly accepted meanings of the words of my title, but also of their dictionary definitions. In Webster's dictionary you will find that "science" comes from the Latin word meaning "to know" and its definitions is "accumulated and accepted knowledge, as of principles or facts, systemized and formulated with reference to the discovery of general truths or the operation of general laws." The word "politics" comes from the Greek word meaning "citizen." Webster gives its meaning as "the art of government; the theory or practice of managing affairs of public policy or political parties."

If we accept these definitions (and Webster is hard to improve upon) then our bedfellows are not so incompatible. Indeed, one wonders why it has taken them so long to court each other. Actually they are not incompatible—they have lived together in separate worlds and just had not been introduced to each other. Never having been introduced, they could not be expected to understand each other.

It took World War II to introduce them. For the first time in the history of the world, there has been a full partnership between the government and the forces of science and technology. This partnership revolutionized military strategy and problems of national defense.

Full text of an address presented at the National Electronics Conference, Chicago, Ill., October 3–5, 1946.

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The fruits of this partnership now are known to everyone. They include radar, rockets, jet-propelled aircraft, guided missiles, proximity fuses, and, of course, the acme of all destructive weapons, the atomic bomb. In all of these, your science of electronics is concerned. Further knowledge in the fields of chemical warfare and bacteriological warfare is certainly hidden in our laboratories. The war through which we have just passed was indeed a total war, for even the psychologist and the economist played important parts and put their knowledge to use in the conflict. We saw the results of carefully planned and well-organized operations in both psychological and economic warfare. Fortunately, almost all the scientific developments of war are useful for the peace. The same knowledge that can be used so effectively to destroy can be used also to build.

FRUITS OF THE UNION

Even in peace, man is constantly at war. There is the urgent and ceaseless war against disease, both physical and mental; the war against the insect world; the war against the forces of nature; the battle against hunger. We fight constantly against our social sicknesses of racial discrimination and disrespect for law and order, and against ignorance and the abuse of power. Our success in the constant warfare of peace depends, in my opinion, primarily upon how well we utilize our scientific resources—how well the partnership of knowledge and man, or science and politics, is joined in the battle.

The men of science and the knowledge gained through their researches have not been partners or participants in the art of government. That art traditionally, and quite properly, belongs to the politician. It should continue to belong to the politicians, for they are essential in any government, but they should be strengthened by the scientists and the knowledge gained by their methods.

The scientist has chosen to be a specialist. His work is so demanding that he cannot assume formal political and social responsiblities. The demands of the laboratory and the nature of his work have made the scientist somewhat inaccessible. Furthermore, there has been a lack of understanding of scientists and their work, as well as a patronizing attitude toward men of science on the part of many men of affairs, an attitude that has prevented the union of these two groups and thus has prevented our society from receiving the full benefit of their joint talents.

The achievements of science under the awful pressure of war have increased the public awareness of the resources of science and its enormous potentiality for the betterment of our lives. The explosion over Hiroshima blasted everyone—scientists, politicians, and plain citizens—into thinking about science and its implications for the future of the world, not only in terms of better things, but also in terms of a better political, social, and economic world.

During the past year, no citizen has thought harder or worried more about the implications of science and technology than has the scientist himself. He, therefore, emerged from the laboratory into the political world to try to explain to the politician what science is, its conditions for healthy growth, and how it should be supported and utilized for public good rather than for the destruction of our world.

The task of explaining their views to the political public proved to be arduous and exasperating, but enlightening also, to the many scientists who took the plunge. The effectiveness of the scientist's attempt to influence public opinion and thinking cannot yet be assessed. It certainly was less effective than he hoped; however real gains were made in the formulation of national policy. One thing is clear, however—the scientist must continue to tell the public what he thinks about both national and international problems facing the nation. During the 1945–46 Congressional sessions, we saw the United States Congress consider and debate two major bills relating to federal support and control of scientific research; the Kilgore–Magnuson Bill establishing a National Science Foundation, and the McMahon Bill establishing federal control of atomic energy. The latter became law, and the former was passed over without action. In addition, at least five other bills concerning government support of science were thrown into the hopper; some for political purposes, some for apparently no reason at all. Even Mrs. Clare Boothe Luce tossed in a bill.

Behind the scenes in Washington, the political and legislative struggle went on for a year. It was a fruitful struggle, not in terms of actual results, but because it introduced the politician and the scientist to each other's brand of thinking, to each other's problems, with each striving to understand the other. The method used by the scientist in his thinking is a process of reasoning *from* facts to conclusions which fit all the facts. Such thinking often leads to new and unexpected conclusions. On the other hand, the politician reasons *from* conclusions into which he attempts to fit all the facts *if* he can. It takes inspired political thinking to produce a result which fits all the facts!

Although the United States Government has supported certain types of scientific research, particularly in the agricultural field, for many years, the issue as to how far it should go in the support of a truly broad and national scientific program still remains unsettled. Fundamental research, upon which all applied research and engineering depends, has been left largely to individual enterprise on the part of men of genius who are wise enough to know the long-term value of knowledge. The question now is, "Can scientific work be left to happenstance, to the vagaries of a vanishing philanthropy, to universities with slim research budgets, to an industry demanding immediate returns on any investment?"

The answer is, "No," provided that a way can be found for the government to support science without restricting the freedom of investigation and the exchange of knowledge, a freedom without which science cannot thrive.

A NATIONAL SCIENCE PROGRAM

There are several compelling reasons why our government must subsidize and support a national science program and not leave the matter to chance.

Our democratic system of government and our belief in private enterprise are grounded in our happy experiences with them, but that experience is only 150 years old. Throughout this brief span we have been operating in an expanding economy with new frontiers of land, increasing population, and new resources constantly opening for us. These frontiers, as we once knew them, now have vanished and our population is rapidly approaching a state of equilibrium. The wise man, who loves his way of life, asks whether his system can endure and flourish under the changed conditions he sees ahead.

While frontiers in the pioneer sense of the word are gone, the new and truly endless frontier of science always will be ahead. It contains vast riches for those wise enough to find and to train the explorers, and to equip the necessary expeditions. I have no doubt that science can guarantee its part in the development of a new golden age for our people, but I doubt and fear lest the people of the world and their leaders are not wise enough to govern themselves.

We are living in a world of cruel reality and the issue of national security is constantly with us. National defense no longer is a matter of the size of the Army, Navy, and Air Force. It is a matter directly related to the total strength and resources of the nation, of which the scientific laboratories are as vital as any other part. That nation whose scientific stature and health are the greatest is thereby the strongest and most secure. Finally, it is true, almost without exception, that the public eventually receives the benefits of any new knowledge. The public, therefore, has a direct interest in supporting research, just as it supports education on all levels, for the public is the final recipient. Both industry and the universities will continue to support scientific research—the former because it pays dividends, and the latter because they have been given the function of training scientists. Neither industrial laboratories nor university laboratories nor federal laboratories can do the whole job. Each has its place in the total scientific picture. There are vast areas of research in nutrition, psychiatry, psychology, chemistry, physics, sociology, and economics, of such a long-range nature and of such complexity that only the government can make the required investment, the returns from which will come far in the future.

The establishment of a sound federal policy for the promotion of science is a matter of direct concern to everyone here today, and each of you must do your own thinking about it—more than that, you must act after you have thought.

Any federal policy or any legislation that may come in the future must result from the scientist and the politician working together. It must satisfy the scientists' demand for high scientific standards, not only in theory, but in practice. It must be sufficiently inclusive not to leave any essential group unrepresented. It must contain provisions acceptable to the legislative and executive branches of the government, since all legislation must be so designed as to work in the political sense. To make our Congress adopt a fruitful policy and pass good legislation depends upon successful political methods. All legislators must be informed regarding the merit to be sought and the evils to be avoided. Any policy or legislation must have a vast majority of scientists and educators supporting it. Legislators must be convinced that the federal support of science is worth the cost. The general public must be sympathetic; at least not unsympathetic. The art of government has its own methods of accomplishing its work. The scientist can do only what any other citizen can do with political methods—study them and follow them until the job is done.

I close with a warning. When our government, or any other government, supports something, it exacts a measure of control. There is no such thing as controlled science, for when science is controlled it no longer is science. We have just seen a nation of great and talented people destroyed in Germany, which, under Hitler, abolished intellectual freedom and passed the control of scientific thinking into the hands of a government of venal and foolish politicians. It can happen here—unless all scientists become citizen-politicians. Making science and politics work together for the national welfare is your job now.

Sliding Mercury Contact. It always has been difficult to transfer high currents from a stationary to a moving object. Machinery requiring large currents has used heavy flexible leads, flexible copper laminations, or large numbers of carbon brushes, and has had a limited length of motion for moving current-carrying parts. A high current capacity sliding contact developed to improve these conditions is described by M. B. Austin of Coatesville, Pa., in the *Iron and Steel Engineer*, September 1946. It consists of a copper container holding a mercury bath in which is suspended, so as to be separated from the container, an upper copper contact, the contact surface of which is covered completely by mercury. Either the container, upper contact, or both may move, motion between them causing both contacts to slide on a mercury film one eighth of an inch thick. Temperature rise in this type of contact is 15 degrees Fahrenheit at a current density of 500 amperes per square inch per 1/64 inch length of current path through the mercury, and 24 degrees Fahrenheit at a current density of 750 amperes per square inch under the same condition, both rises occurring during a 5-minute period of contact operation.

Slow-Acting Relays

I. Historical Aspects

ARTHUR BESSEY SMITH
FELLOW AIEE

Slow-acting relays have been the unobtrusive little behind-the-scenes timing-control elements in various kinds of electrical installations for so long that they now are taken for granted. Although supervisory control, telemetering, and protective relay systems for electric power systems, have been covered extensively in technical literature, the role played by the slow-acting relay for the most part has been passed over. In a like manner, much has been written about machine switching in automatic telephony, but little has been said about the slow-acting relays at the heart of the works. Aware of this deficiency in published information, the AIEE committee on communications, under direct sponsorship of the subcommittee on quick and slow-acting relays, arranged a conference at the 1946 AIEE winter convention for the purpose of preliminary exploration of the subject and of the interest in it. The essential substance of that material is presented in these articles.

IT IS NECESSARY to understand clearly what is meant by a relay which is slow to operate or slow to release. It is not a matter of definite time, such as so many milliseconds. It is relative to the speed of the great mass of relays used in the particular field under consideration. In a field where most relays operate in 5 or 10 milliseconds, one which intentionally takes as much as 30 milliseconds to operate is clearly a slow-operating relay. But in a field in which the great run of relays take 40 to 70 milliseconds to operate, a relay would need to delay its operation as much as 150 to 200 milliseconds to classify as slow to operate.

Today slow relays are used in a number of fields, such as electric power, telegraph, and telephone. Most of these fields began to use slow relays in the era from about 1900 to 1910.

ELECTRIC POWER

At least as early as 1904 there was in use a relay of the solenoid and plunger type, fitted with a leather bellows having a small hole in one head. The slow escape of the air delayed the completion of the stroke of the plunger to the desired amount. Some relays of this type are still in use.

Later relays of the induction motor type were used, in which the delay was obtained by means of a disk. Inertia played a part in the delay and sometimes small permanent magnets were added, to cause eddy currents in the disk, similar to the structure in the integrating watt-hour meter.

TELEGRAPH

Hand-operated Morse telegraph sets seemed to need no slow relays. The circuit must be under just as perfect control as is practical, so that the operator can make dots, spaces, and dashes, long and short, at will.

Before 1909 a ticker system used by the New York Quotation Company employed a "time relay." It had a long stroke to obtain a desired delay in operation.

In the same era there was a step-by-step telegraph printer which had a "press magnet" whose armature was heavy enough to furnish considerable inertia. During the flow of rapidly alternating impulses the armature could not respond. But on steady direct current it operated and printed the character. The same device was used in the Gold stock ticker and in the Essick page and line printer.

TELEPHONE

It was in the telephone field that relays fully came to their own, as to numbers and importance to the art. Because such large numbers of relays are needed, they must be relatively cheap. Because of limited space they cannot have the desirable cube form, but must be relatively long and slender. This enables the maintainer to get at any relay at either end without taking it off the shelf or rack. Early manual switchboards had no need for slow relays. In fact, speed was not considered.

The first automatic switches were operated by the hand of the subscriber, using push buttons or strap keys, with so many pushes for each digit. Time was not much of a consideration, for the two wires of the subscriber line were grounded in turn in accordance with a definite code. Numerical impulses were sent over one line wire and one impulse over the other wire operated the "numerical separator" so that the next group of impulses on the first wire would operate a different magnet. There was no need for slow relays.

The entrance of a definitely slow-to-release relay is interesting.

In 1900 the New Bedford, Mass., automatic telephone exchange was installed. Each selector had two series line relays, each 30 ohms in resistance. These remained in series with the line wires after the connection was extended to the next switch. Thus the completed con-

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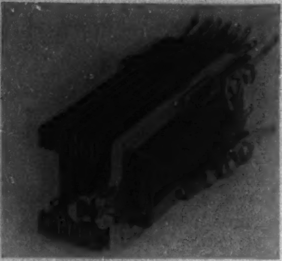

Figure 1. The *U*-type relay

nection had several pairs of 30-ohm relays in series, through which the subscribers must talk (local battery). To improve the voice transmission, each relay was shunted with a 150-ohm noninductive resistance.

In 1901 the Fall River, Mass., exchange was installed. It had the same 30-ohm series relays in each switch, but the iron core of each relay was covered by a copper tube, whose wall was about 0.80 millimeter thick. The same copper sleeve was applied to several magnets used in lifting and rotating the switch-wiper shaft. In the latter case it materially reduced the spark produced at the contact of the relay which controlled such magnets.

In the same plant, and at about the same time, it was discovered that short-circuited turns in a relay winding, as well as the copper sleeve on the core, had a strong delaying effect on the release of the relay. In fact, it was found that if the copper sleeve were thick enough, it tended to run the impulses together.

At least as early as 1904 there were a number of a-c relays in use. The armature, instead of being pivoted at one end, was pivoted somewhere near the middle, and it was capable of rotating as much as 40 or 45 degrees before closing the contact spring. This had in it great possibilities as a slow d-c relay, though it seems not to have been used that way.

In 1904, during the month of July, an engineer of the Automatic Electric Company devised the first 2-wire common battery subscriber line circuit, combining quick and slow relays in numerical selection. At the subscriber station there was no ground connection (except for the lightning arrester). When the subscriber seized the line by closing the loop, a quick line relay operated and pulled up a slow relay. The latter grounded the third wire and prepared local circuits for receiving numerical impulses. As the subscriber dialed the first digit, the line was opened several times, each opening causing the operation of the motor magnet which moved the switch wipers. During this time the slow relay remained operated. A second slow relay was associated with the motor magnet—this relay operated and remained operated until the end of the group of impulses. Then it released and prepared the local circuits for the next group of impulses (the second digit of the call number). These slow-to-release relays were made so by copper sleeves.

In 1908 the Kellogg Switchboard and Supply Company needed a trunk circuit between an automatic exchange and one of their own manual switchboards. The company devised a circuit for the manual end of the trunk line which would ground both the trunk wires, then clear both from earth. This was done by the use of two quick relays and one slow-release relay. When the operator pulled down the connection, the first quick relay let go and connected the two trunk wires to earth, also through the other quick relay. The latter operated on the release current, and cut off the slow relay, which soon released and cleared the lines.

By 1910 the slow relay had been changed by its use of a copper collar instead of the copper sleeve. By placing the copper collar on either end of the relay core, great differences in performance can be obtained. Because of the slender shape of the electromagnet, if the collar is at the armature end of the core, the delay in operating time may be as much as 75 to 100 milliseconds. If it is at the "heel" end of the core, the delay to operate may be only 10 to 15 milliseconds. This difference is a result of the magnetic leakage between core and heelpiece, which would not be very effective if the magnet had a cubic shape.

In 1930 there was a return to the sleeve for the release and magnet relays, in order to diminish the cost. The cost of the copper and its labor was reduced about 40 per cent and that of the complete coil from 10 to 15 per cent. There has been no impairment in the performance, but the copper collar is retained wherever its specific performance is of value.

Other means, such as weighted springs or clockworks with air for governors, may be used for slowing the relay but only the copper collar and sleeve seem to have obtained any significant degree of usage in the telephone field.

II. Use in Telephone Switching

H. N. WAGAR

A TYPICAL RELAY, used by the millions in telephone offices, in shown in Figure 1—the *U*-type relay. This particular type is not a slow relay; its action time is about 1/20th second or less. It can be slowed, however, to give times in the range from 50 to 500 milliseconds, which is the ordinary range of delay in telephone relays. These delays can be obtained—at reasonable cost, with reasonable accuracy, and without too great a sacrifice in number of contacts—through application of the principles to be outlined. Primarily,

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these delays are obtained by the use of short-circuited turns to oppose a change in magnetization, and by using the properties of the relay upon disconnection of the circuit, not on closure, because of the greater stability of a relay upon its release. The subject of this article then narrows down to the principles underlying slow-release relays.

At the bottom of Figure 2 appears the "slow-release formula," by which almost any slow-release problem may be explained. Because of its importance, the steps by which this formula is reached will be outlined. At the top of the figure is a simplified relay structure. When the switch *S* is closed, the armature is attracted from its unoperated position to one against the core. The condition of magnetization is represented by the dotted φ versus Ni curve at the right, the point corresponding to φ_1 representing conditions after the armature has come to rest against the core. In this operated position, the armature experiences a mechanical force, *F*, resulting from the deflected springs trying to restore it to normal, and a magnetic force, *P*, holding it against the core. This magnetic force depends on the flux in the air gap, and the pole face area, *S*, according to the formula

$$P=\frac{\varphi^2}{8\pi 980S}\text{(in grams)}$$

When the switch is opened, the applied ampere-turns eventually will drop to zero, and the flux follows according to its demagnetization characteristic. With the flux, of course, the pull also falls off.

The release time, therefore, will be the time, after disconnecting the switch, at which equilibrium between magnetic and mechanical forces is just unbalanced. It is, then, the time for the flux to fall from its steady value to the value of flux at which the pull, *P*, resulting from magnetic attraction equals the spring force, *F*. Looking at the heavy curve to the right, it depends on the time for the flux to travel its downward φ-Ni characteristic (its demagnetization curve) to the just-releasing point. This time would be negligible were it not for the eddy-currents induced in the second winding shown, a short-circuited winding which opposes any change in flux linking its turns. All this may be expressed mathematically by writing down the voltages around the circuit: the *IR* drop plus the back electromotive force, which results from a changing flux, of $N(d\varphi/dt)$. This is shown written in correct units, and when terms are rearranged gives the slow-release formula already mentioned.

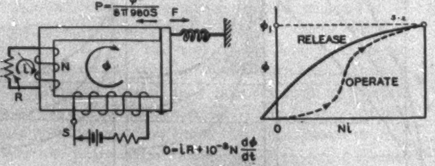

Figure 2. Slow release relay

Slow release formula:

$$t=10^{-8}\frac{N^2}{R}\int^{\varphi_1}\frac{d\varphi}{Ni}\text{ second}$$

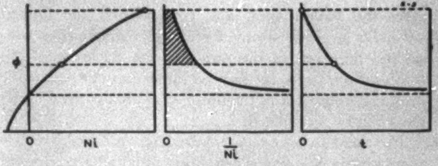

Figure 3. Slow release formula split into conductive and inductive terms

Slow release formula = *(conductive term)* × *(inductive term)*

When this formula is used, it is convenient to split it into two separate terms, as shown in Figure 3. The N^2/R term has dimensions of a conductance and is called the "conductive term"; it depends solely on the dimensions of the coil and the quality of the conductor. The remainder of the expression is called the "inductive term":

$$10^{-8}\int^{\varphi_1}\frac{d\varphi}{Ni}$$

It is a function of the iron quality and the iron dimensions. In other words, it depends on the descending φ–Ni relation of the magnet, a relation amenable to both measurement and prediction. By replotting the flux against $1/Ni$, the information is arranged in the form $\int xdy$, which expresses the area included between the curve, the vertical axis, and the two ordinate limits. This area is shown shaded in Figure 3. The product of this area and the conductive term gives the desired time in seconds, varying as shown at the extreme right of the figure. A more detailed examination of each of these terms will be made next.

The manner in which the conductive term varies is seen in Figure 4, where the conductance of a coil (in megamhos per inch of length) is plotted against the ratio of outside to inside radii. The heavy curve, representing a solid sleeve, is seen to give the greatest delay in a given space, as might be expected. The other curves show the results for the extremes of typical windings where the value of e represents the "copper efficiency" of magnet wire; that is, the ratio of copper volume to total volume.

The inductive term in Figure 5 can be used to estimate effects of iron, gap, or other dimensions. As an illustration, the three curves may be considered as those for three different operated gap-lengths, differing only by a few tenths of a mil from each other. The replot of these values against $1/Ni$ shows how the smallest gap gives the largest area, and consequently the longest time, for a given spring force. The dimensions considered here are in the order of the thickness of the finish on the parts, so that the important effect on time resulting from wear of the finish or manufacturing variations can be recognized.

The three cases being compared here are seen to differ widely in time upon the assumption that all are adjusted to the same operated load. Should it be possible to change the loads so that in each of the three cases the just-release ampere turns were the same, it will be evident by reading across from the set of curves on the left side of Figure 5 to that on the right, that the releasing time will be approximately the same for each case. This relationship between just-release ampere turns and time is the basis for large-scale manufacture of the slow-release telephone relay known as the *Y*-type relay to be described further in the following. So long as springs are provided to give the needed releasing load, the time is known when the release *Ni* are known.

The curves also could be used to analyze the effectiveness of various pole face areas. In general, a large pole face will be found best suited to long delays, but an optimum dimension sometimes will be indicated by this type of analysis. The use of these methods gives the designer reliable results. Figure 6 shows agreement between prediction and experiment in a particular test sample; where the heavy curve shows the calculated time plotted against releasing *Ni*, and the circles show measured points.

To summarize, then, the design features of the slow release relays are as follows:

1. Short-circuited windings.
2. Low-reluctance gaps.
3. Low-leakage magnetic design.
4. Saturated iron (high current).
5. Low spring forces.
6. Massive moving parts.

The designer will build his slow relay using a solid sleeve of copper wherever possible.

The smallest possible operated gaps consistent with corrosion and wear resistance will be used, and a large pole face area will be helpful. An efficient magnetic circuit, operated into the saturated region, will be best. Light restoring forces will require a longer flux decay time to reach. For completeness, massiveness of the moving parts is mentioned, though it usually is of secondary importance.

Such principles are embodied in the *Y*-type relay shown in Figure 7. This relay uses copper sleeves, iron saturated by 200 to 250 ampere turns, a low-leakage magnetic circuit, nickel finish for corrosion protection, and a thin chromium finish for wear protection. While it uses nearly iron-to-iron contact in the operated position, an important feature is a slight sacrifice in operated gap reluctance to obtain greater stability—the spherical gap surface (visible in the figure), which gives uniformity at the gap in spite of misalignments encountered with normal manufacturing practice. Its other feature, as previously mentioned, is its ability to be adjusted for a given time requirement simply by adjusting to release on a specified value of ampere turns. The extra springs provided to insure that the releasing force can be attained in all cases will be seen in the figure.

III. Telephone-Type Relays

ANDREW W. VINCENT

THE TIME INTERVAL required for the closing or opening of a relay contact after the current is applied or removed, is dependent upon the balance of mechanical forces acting on the mass of the armature. These balanced forces are the spring load force, the magnetic operating force, the frictional force, and the force of acceleration. Because changes in the magnetic forces occur when the armature is unoperated, while it is moving through its stroke, and also after it is operated, it is necessary to show the relationship of the magnetic forces as a function of armature stroke at constant ampere-turn values and the magnetic forces as a func-

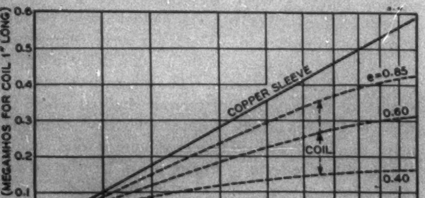

Figure 4. The conductive term N^2/R

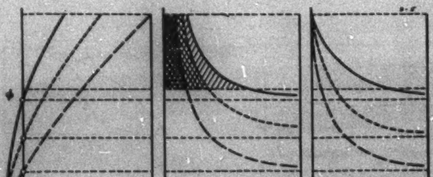

Figure 5. The inductive term $10^{-8}\int_{\varphi}^{\varphi_1}\frac{d\varphi}{Ni}$

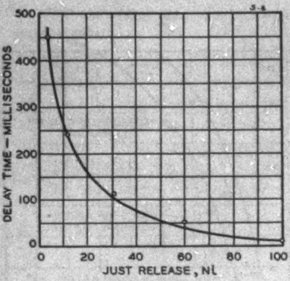

Figure 6. Agreement between prediction and experiment in a particular test sample

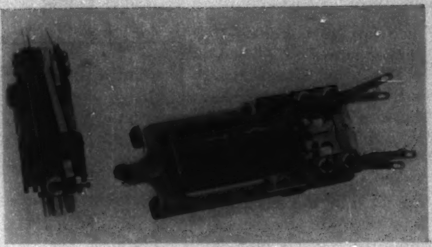

Figure 7. The *Y*-type relay

tion of ampere turns at various armature stroke positions. The curve of Figure 8 shows these relationships in a typical telephone-type relay structure carrying an operating load of 12 make-contacts and operating with zero residual air gap.

The left half of the curve sheet shows the magnetic force versus armature stroke at different ampere-turn values, and the spring load force versus stroke of the 12 make-contact spring load. The dotted curves show the build-up and build-down forces under typical operation. The right half of the curve sheet shows the build-up of force at operated and unoperated stroke positions after demagnetization and build-down after 600 ampere-turn saturation. Starting at 0 and 0′, and progressing through the numbered points back to zero, shows the static forces throughout the armature stroke. The following is a detailed discussion of the cycle.

From 0′ to 1′ represents the build-up of the magnetic force in the unoperated armature position to the point where it is greater than the spring load force at 1, at which time armature motion begins. The time required for this part of the cycle is dependent upon the time required for the current in the coil to build up to the required value as given by

$$i=(E/R)(1-\epsilon^{-Rt/L})$$

where L equals the inductance of the coil with the armature in unoperated position.

When the eddy current losses in the iron are large, the accompanying counter flux may cause the time to be much longer than the foregoing equation would indicate. Advantage has been taken of this in adding short-circuited turns to the winding to increase the operating time. The short circuit may take the form of a relay winding or a solid copper collar. The location of the short-circuiting winding must be kept close to the armature to prevent the leakage flux from acting on the armature. Inasmuch as the armature is open, there is more of a tendency for the flux to take the leakage path.

Points 1 to 2 indicate the rise of force as the armature closes. Although the spring load would be lifted by 110 ampere turns as indicated by the fact that the spring load is less than the 110 ampere-turn force curve throughout the stroke, it is necessary to saturate the structure with 600 ampere turns in order to show the most severe release condition. When the armature has completed its stroke, the ampere turns have increased to 150 which is shown at 2. The contacts close at the point indicated, the elapsed time being that necessary for the changing force, represented by the difference between the spring load force and the dotted force curve, to move the effective mass of the armature and springs to the point where the contacts close. As the current has not reached its final state, the force increases at a constant closed gap from 2′ to 3′ which is outside the curve sheet. The points 2 and 2′ would not coincide exactly because of the different paths taken, but the difference is assumed to be negligible. Curve 0′2′3′ is similar to the normal magnetization curve of the iron, and the upper curve is similar to the upper part of the hysteresis loop, except that the force is a squared function of the flux. The equation is

$$F=\frac{\phi^2}{7{,}840\pi A}\text{ grams}$$

when

ϕ = flux in maxwells
A = area of the gap in square centimeters

When the relay coil circuit is opened, the force decreases along the curve 3′ to 4′ until the magnetic force is less than the spring load, at which time armature motion begins. The elapsed time is again dependent upon the effectiveness of the eddy current paths in building up currents resulting from the changing flux which builds up counter flux to oppose the change in flux and thus maintain the force. Inasmuch as the armature is closed, the position of the short-circuiting conductor is not as important as in the case of the unoperated armature position. The rate at which the circuit contact is opened also affects the time, especially if there are no short-circuited turns, since the large change in flux builds up high voltages which cause current to flow as an arc.

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The point 4′ indicates that the relay will release on 35 ampere turns.

The dotted release force curve 4 to 0 indicates the decrease in magnetic force as the armature opens. The force decreases rapidly as the armature is opened inasmuch as the flux path is broken. The time elapsed until the contacts open is dependent upon the applied forces and the mass of the system as in the operating case.

In practice, the determination of relay release and operating times is limited because of the large number of variables.

The value of the voltage applied to the coil is not an important factor in the consideration of release time because the magnetic structure is near saturation with the armature closed, under average operating conditions. The release time is largely dependent upon the rate at which the flux decays. The time required for armature motion can be neglected in slow releasing relays as it is a small part of the total release time. By knowing the time required for the force to drop to the operated spring load force at various residual air gaps and with a specified short-circuiting collar, the release time can be determined. With a large copper sleeve on the core release time as high as one-half second can be achieved without failure to release.

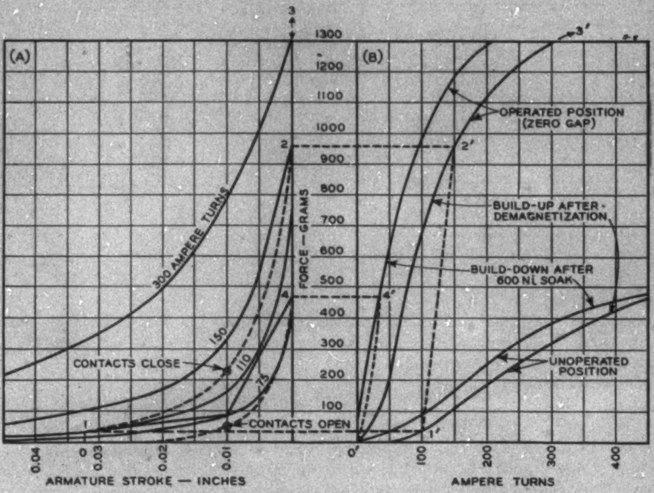

Figure 8. Relationship of magnetic forces as a function of:

A—Force versus stroke at constant ampere turns B—Force versus ampere turns at constant stroke

The decrease in height of the residual projection, resulting from operation wear, causes a greater increase in release time when smaller residual setting adjustment is used, inasmuch as the change resulting from wear is a greater percentage of the total residual air gap. When the effect of magnetic iron aging, which may double the residual force over a period of time, is taken into account along with the accompanying decrease in spring load force, it can be seen that the release time increases with the number of operations, approaching the point where the armature will fail to release.

In determining the operate time the applied voltage cannot be neglected. Over the normal operating ampere-turns range the introduction of the unoperated armature air gap into the magnetic circuit shifts the operating range of the iron to a lower region such that current changes are not masked by saturation. For this reason the applied voltage must be considered inasmuch as it, with the L/R ratio of the coil, determines the current in the coil as a function of time. The inductance of the coil is less with the unoperated armature air gap which is responsible for a maximum practical limit which appears to be about one-tenth second operating time delay.

IV. Use for Power Purposes

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FELLOW AIEE

THE APPLICATION of telephone relays in the power field is not new. These relays have been in use for years by many utility companies on special applications. The larger power manufacturers have used them as elemental units in telemetering, selective calling on carrier telephone circuits, carrier relay protection, protective distance relaying, and also in supervisory control.

Extensive use has been made of standard telephone relays in the regular, slow operate, and slow release types. In addition to the standard cylindrical-type relays the following relays or relay devices have been used by some power utilities:

1. Rotary selector switches.
2. Morkrum cable relays.
3. Double throw polarized relays.
4. Weighted reed timing relays.

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5. Toll line ringing assemblies for 20-cycle ringing and for 135-cycle ringing.

The major use of such relays is as auxiliary devices, but they also are used a great deal for automatic reclosing, carrier signaling, private telephone line ringing, timing, sequence control, interlocking, and time delay purposes. The characteristics of standard relays are visualized easily if plotted on log-log paper.

The only major disadvantage of telephone relays is in their mounting requirements but this largely has been overcome by the design of a special front-of-board mounting case with studs for rear connection.

The advantages of these relays are their low price, small space requirements, wide range of contact assemblies, resistance range from $1^{1}/_{2}$ to 15,000 ohms, reasonably accurate calibration, dependable design and test procedure, adequate engineering data, good delivery (normally), and low energy requirements.

Organization of the Engineering Profession

ELGIN B. ROBERTSON
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This searching consideration of the question of organizing or reorganizing the engineering profession is presented for the purpose of stimulating additional thought and discussion by the entire AIEE membership. The author, speaking in the light of experience as president of the Texas Society of Professional Engineers and vice-chairman of the AIEE membership committee, supports Plan B of the various proposals of the AIEE planning and co-ordination committee's professional activities subcommittee (EE, Apr '46, pp 169–73). He urges that collaboration and co-operation are more effective than attempted consolidation in view of the great diversity of interests among engineers.

IN CONSIDERING the organization of the engineering profession, there are two distinct phases which must receive attention—the *technical* and the *professional*.

The technical side of our profession is well organized. We have the Founder Societies and a host of other societies, each devoting itself principally to the dissemination of knowledge to its members. Our AIEE is more or less typical of these technical societies. It is certainly a going concern. It is the largest of this group. It receives a continuing infusion of new ideas through the new members, and the new local and national officers elected periodically. Certainly, as an organization, AIEE does not need any particular "going over." Any radical change in its organization, or of its set up, would not seem to be warranted, and I doubt that such changes would meet the approval of our membership. All these characteristics are but indications of a successful organization.

Let us not lose sight of the fact that our AIEE and other major technical societies are, as a rule, old organizations representing a wealth of accumulated experience and well-earned prestige. Let us remember that the success they now enjoy is the result of the thought, the time, the energy, and the money put into them by their members over a period of a half-century or more. Major changes in either status or major objectives should be approached with great care.

When we consider the professional side we find an entirely different situation, and it is with this phase that I wish to deal mainly.

What is a profession? One answer to this question can be gotten by referring to "Webster," but one will find only the definition of the *word*. Surely our engineering profession is more than can be defined by a single word, no matter how comprehensive that definition may be. But, regardless of any definition, I believe that a profession is what its members want it to be; inevitably, it is whatever they make it. If we want our engineering profession to be strong, virile, and honorable, then we, its members, must breathe into it the desired life, vitality, and honor through the integration of our individual actions. If we do not want this kind of a profession, then we can refuse to give to it our time, our energy, and our money. If we are

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personally selfish or thoughtless we may conduct ourselves so as to get the maximum benefit for ourselves, regardless of the effect such conduct may have on the welfare of the whole group. Some of us are doing just this, although the vast majority are not.

Every engineer has his technical needs and his professional needs. Both of these concern his welfare and so he should devote himself to both of these activities.

I have been told time and time again that engineers cannot divide their loyalty between technical and professional societies and that they do not have enough spare time to devote themselves to both. I have been associated with AIEE for approximately 30 years. I started out as a Student Member and have advanced through the various grades to that of Fellow. I have been a Section secretary and Section chairman, have served on numerous committees, and as a vice-chairman of the membership committee for the South West District. I have belonged to the Texas Society of Professional Engineers, and the National Society of Professional Engineers for several years. I have been a chapter secretary, a chapter president, a state director, and in January of this year, I completed my term of office as president of the Texas Society of Professional Engineers. Has this activity in the professional society lessened my interest in the AIEE? By no means! As a matter of fact, it has enhanced my interest in and appreciation of AIEE. This division of activity is a manifestation of breadth rather than division of professional interest. You cannot divide your loyalty to yourself; nor can you give to your societies (technical or professional) *only* time that you consider to be *spare* time if you wish them to prosper and to achieve the results you desire.

A technical society such as the AIEE cannot be an all-inclusive society. The technical problems of the electrical engineers are not the same as the problems of the civil, petroleum, or other engineers. However, a professional society can be all-inclusive because the professional problems of the electrical engineers are identical with those of all other engineers. These professional problems are not limited to the engineers in private practice but are the concern of all engineers.

Why are we interested in organizing the engineering profession and why are we discussing these professional problems? Actually, if we will be honest, we must admit that this interest is aroused by our desire to have the public give us the recognition, as a profession, to which we think we are entitled.

All right—we admit that we want public recognition and esteem as a profession. How will we get it?

Who is an engineer? Answer this question by taking a piece of paper and writing down your own definition of an engineer. If you will take the definitions given by ten different engineers, you will have ten different answers.

Who is a doctor? He is one who is licensed to practice medicine by the laws of his state. A doctor of medicine degree from a medical school, or years of experience, do not entitle him to practice. It is the license issued by his state that is the mark of identification and authorization. Who is a lawyer? He is one who is licensed to practice law by the laws of his state.

By the same token, I believe that we should define an engineer as one who is licensed to practice engineering by the laws of his state. If we of the profession cannot put a label on the engineer, then how can we expect the public to recognize an engineer when it sees one or hears one? We must be able to point definitely to a particular individual and say: "This is an engineer." I believe that the legal license is the best answer to this simple identification.

I believe that most of us will admit that the lawyers have influence in their communities. I believe that we will concede further that the public is conscious of the professional status of the doctor. Would either of these be in their present position without the Bar Association and the Medical Society? A professional engineering society is essential to professional recognition and standing.

The professional activities subcommittee of the AIEE committee on planning and co-ordination has made a report on four plans for organizing the engineering profession (*EE, Apr '46, pp 169–73*). This report is worthy of every engineer's thought and consideration. I believe that the course presented by "Plan *B*" of that report is the best to follow. Plan *B* leaves the AIEE and other technical societies to carry on as they now do and calls for a professional society to carry on all activities that are of a nontechnical nature. I believe that Plan *B* will do the job we want done.

As the all-inclusive professional society called for in Plan *B*, I personally favor the National Society of Professional Engineers because

1. It is already in existence as a going concern and ready to serve.
2. A license is a prerequisite to its membership.
3. It is a thoroughly democratic organization. Its members are elected locally. The state societies and the national society cannot elect a member. A state society is primarily a co-ordinating agency of the local chapters in that state and the same is true of the national society in its relation to the state societies.

This professional problem is primarily of local concern and since all power in this organization originates with the individual member and his local chapter, it is particularly effective in handling the questions that arise. The local chapters can work through the state and national societies in taking care of those problems that are more than local in scope.

This organization frankly says that it is concerned with the welfare of the individual engineer, and since this is the main concern of all I think it is worthy of the support of all engineers who want an engineering profession that is entitled to the esteem and respect of the public.

Great Lakes District Meeting Conference Papers Digested

Conferences on subjects varying from electric machinery to basic sciences were held at the Great Lakes District meeting of the Institute held at Indianapolis, Ind., from October 9–11. The majority of the papers presented were of the conference or District type and will not be published formally. The following material summarizes the papers presented at the meeting.

CONFERENCE ON ELECTRIC MACHINERY

The conference on electric machinery was presided over by Chairman M. S. Oldacre (M '42) Commonwealth Edison Company, Chicago, Ill. The conference covered a wide range of material from theoretical design to present-day commercial standardization of small motors. Outstanding features of the meeting were the papers on magnet wire and fractional-horsepower motor standardization. The paper by C. P. Potter on standardization stimulated the greatest amount of discussion, and it seemed clear that adoption of these standards will be of great benefit to both users and manufacturers in this $100,000,000 a year industry. (*EE, Nov '46, p 451*)

"Asynchronous and Single Phase Operation of Synchronous Machines" by A. W. Rankin (M '45) General Electric Company, Schenectady, N. Y. This is the fourth in a series of analytical studies of synchronous machines made by Mr. Rankin, the other three already having been published in the 1945 AIEE *Transactions*. In this current paper, Mr. Rankin presents an analysis of the asynchronous and single phase operation of a synchronous machine based upon the generalized equivalent circuit of Kron. The fundamental equations of Park are used as the starting point, but the succeeding mathematical operations are guided by the tensor publications of Kron. The asynchronous torque and current relations are presented in terms of the currents and voltages of simple equivalent circuits; also, a numerical comparison is given of the asynchronous and single phase operating characteristics of synchronous machines with complete and with incomplete end rings on the damper windings. This comparison, the author states, has direct value in design engineering, not only as a study of the various torques, but also as a picture of the distribution of the damper-bar currents in the leading and the trailing pole halves. The methods presented are intended for use with the a-c network analyzer, and all torque and current expressions are given directly in terms of current and voltage readings obtained from the network analyzer. Equivalent circuits for constant-speed asynchronous operation are derived, and numerical examples of the application of these equivalent circuits are given. Equivalent circuits for constant-speed single-phase operation also are derived, and a single numerical example of the application of these equivalent circuits is given.

"Fractional-Horsepower Motors for Operation on Thyratrons" by W. R. Goss and F. T. Carlson (nonmembers) of the General Electric Company, Fort Wayne, Ind. The many special applications of motors to welding-head controls and other such equipment where the motors are operated by power from thyratron rectifiers have led to the development of "packaged" power rectifier equipment by several manufacturers for use in driving variable speed motors on many types of application. The availability of this packaged thyratron-rectifier power unit has increased the tendency to utilize standard motors, a practice which the authors report as being not entirely satisfactory because of the additional heating, poor commutation, and noise, which results from the use of thyratron-rectified power on motors not designed for that kind of a power source. Specially designed and specially rated motors have been developed for use with thyratron-rectified power, thus making the variable speed qualities of a d-c motor fully and readily obtainable wherever a-c power is available, through thyratron rectification.

"Properties of Magnet Wire" by H. A. Smith (A '46) and W. G. Skinner (nonmember) General Electric Company, Fort Wayne, Ind. Magnet wire is one of the most important materials used in the construction of electric apparatus. More than 200 million pounds of copper magnet wire are produced annually in the United States. Although it is not generally realized, the magnet-wire industry is a precision industry. Tolerances which would be close in a precision tool room are not uncommon on small sizes of wire, such as the 0.0001 inch plus or minus that is allowed for wire of the size of an ordinary hair, about 0.003 inch. In the past, the acceptability of magnet wire was determined by trial-and-error methods, the design engineer winding such wire into a model of his apparatus and then testing it to destruction. Now, test methods for the evaluation of magnet wire are used by the wire engineer. An attempt is made to correlate the properties of insulated magnet wire—physical, chemical, and electrical—measured by wire tests to the properties desired in the finished electric apparatus. Of the 200 million pounds of magnet wire used by the electrical industry in a normal year, some 70 per cent is film coated—that is, enameled. The two main types of wire enamel used today are synthetic resin enamels and oleoresinous enamels based upon natural resins in combination with drying oils. Of the several types of synthetic-resin wire enamels now in use, the major one is enamel based upon polyvinyl acetal resin, the properties of which are far superior to those of a natural resin enamel for film-coated wire.

"New NEMA Fractional-Horsepower Motor Standards" by C. P. Potter (F '29) Wagner Electric Corporation, St. Louis, Mo. There are millions of fractional-horsepower motors in operation in the United States at the present time driving washing machines, stokers, oil burners, refrigerators, domestic water pumps, gasoline dispensing pumps and an almost limitless list of other devices including business machines. The public has become accustomed to having these devices operate essentially night and day with very few interruptions. Every fractional-horsepower motor has a horsepower rating stamped on its name plate, and this would seem to be a definite indication of its capabilities. However, this is not necessarily true, because motor characteristics frequently are modified to suit the desires of the manufacturer of the devices which they operate; some modifications being logical, others illogical. These and related practices have resulted in many fractional-horsepower motors not having the life expectancy which they would have if they were more consistently and conservatively designed and applied. To promote a more logical design and application of fractional-horsepower motors, the better to serve the interests of all parties concerned, the National Electric Manufacturers Association recently has adopted two important new fractional-horsepower motor standards: "Definition of Motor Output" and "Service Factor." These prescribed standardized readings, adherence to which would assure to the user a long and dependable service life, would enable the application engineer and the appliance manufacturer to select a dependably standardized motor thus obviating the need to test each application.

CONFERENCE ON ELECTRONICS

The conference on electronics met under the direction of O. W. Livingston (M '43), General Electric Company, Schenectady, N. Y. In summarizing, Livingston stated that a method should be developed by which all vacuum tube characteristics could be expressed in terms of volts, amperes, seconds, and temperature thereby simplifying description and application through standardization. It was interesting to note that the papers on electronics in measurements and aeronautical electronic devices concurred in the necessity for components and vacuum tubes possessing simplicity and ruggedness required in these applications.

"Electronics in Measurements" by R. J. Kryter (A '35) Esterline-Angus Company, Indianapolis, Ind. In spite of the fact that many types of measurements otherwise impossible can be made with the aid of appropriate electronic equipment, electronic methods admittedly have been slow in penetrating the field of instrumentation. Prejudice and sheer human inertia of course have contributed to the slowness of acceptance of electronic instrumentation, but it must be admitted also that it was logical for electronic instruments and measurements methods to be required to demonstrate both their facility and their dependability, and it must be admitted further that there is much room for refinement and improvements in ruggedness and simplicity of the electronic devices used in measurements, particularly in the vacuum tubes themselves. Electronic measurement methods must compete with existing mechanical, electrical, hydraulic, and pneumatic methods, and, initially at least, are most attractive in the fields not so readily served by the more established instrumentation equipment and methods. For example, electronic methods are most advantageously adapted to miscellaneous or special quantities such as light and sound intensities, color values, torque, gas composition, pH, or measurements under special conditions such as speed without physical contact. Similarly, electronic methods are indicated for the measurement of very small quantities, very rapidly changing quantities or conditions, and very short time intervals. Nonelectronic methods commonly require a human operator and manual manipulation, whereas the electronic method and equipment permits continuous and automatic measurement, and automatic control. Extreme sensitivity, great amplification, unlimited speed, smooth and easy control, possibility of automatic operation, are advantages. However, electronic instrumentation is not a cure-all; lack of ruggedness, difficulty of service and requirement for special service personnel, limited life, instability of calibration, and obsolescent components are challenging problems requiring and currently receiving attention.

"Frequency Performance of Thyratrons" by H. H. Wittenberg (A '46) Radio Corporation of America, Victor Division, Lancaster Pa. As there are many applications for thyratrons at frequencies above the standard power frequency of 60 cycles per second—such as in inverters, servomechanism controls, relaxation oscillators, radar modulators, and grid-controlled rectifiers for air-borne equipment, where the frequency range is from a few hundred to as much as 50,000 cycles per second—the performance of thyratrons at these frequencies becomes a matter of some interest and importance. Investigation reveals that at high audio frequencies the grid-controlled characteristic resolves itself into two characteristics—a starting characteristic and an extinguishing characteristic; the starting characteristic shifts negatively with increased frequency, and with increasing gas pressure; the extinguishing characteristics shift negatively with increasing frequencies, with increasing grid resistance, with increasing anode current, and with increasing gas pressure—the shifts also being a function of tube geometry.

"Microwave Magnetrons" by F. F. Reike (nonmember) Purdue University, Lafayette, Ind. In discussing the possible uses of magnetrons—as one of the members of the vacuum tube family—exclusive of radar, Doctor Reike pointed out that the cost of the tube (probably $1,000 each for small quantity production) and the design requirements for the dissipation of heat (about five per cent of rated power from the cathode) were two major problems with which to be contended. He described the steady power output possibilities as ranging from 0.1 to 5 kw, the pulse power or instantaneous output as ranging from 100 to 3,000 kw, with an efficiency of approximately 50 per cent. Aside from radar, the magnetron is considered to have possible applications in the communications field in connection with radio relay links, in industrial heating, and as a source of high power at high frequency for various research purposes. The pulse magnetron is not useful as a continuous-wave source.

"Electron- and Ion-Beam Instruments" by G. W. Dunlap (M '42) General Electric Company, Schenectady, N. Y. Modern beam-type instruments are dependent on a number of requirements common to nearly all. These are: an evacuated space; source of charged particles; means for projecting these particles; beam control; and means for utilizing beam. From the discovery of vacuum technique in 1650, the discovery of charged particles in 1853, and Edison's observation in 1883 of the famous "Edison Effect," the author traced the continuity of fundamental principle and technique in the successive development of various electron-beam and ion-beam devices ranging from the oscillograph to the mass spectrometer to the cyclotron–betatron–synchrotron family of accelerators.

"The Life of Electronic Devices Used by United States Army Air Forces" by A. P. Upton (nonmember) Minneapolis-Honeywell Company, Minneapolis, Minn. Stating that war experience showed conclusively that electron tubes are by far the weakest link in the entire chain of electronic control and other equipment, the author urged that designers and manufacturers take the steps necessary to make industrial electronic components "so rugged and so dependable that they can be soldered into the equipment circuits as easily and assuredly as one can now solder in a resistor or capacitor."

CONFERENCE ON POWER TRANSMISSION AND DISTRIBUTION

The paper on single-pole relaying and reclosing on a 132-kv system attracted considerable attention. It gave the first complete record over a number of years of a new development in tranmission engineering. Problems of rural electrification always seem to produce extended discussion, and this conference was no exception. Technical problems requiring solution relate to production of reliable low-cost devices for sectionalizing lines and to proper co-ordination of fuses and circuit reclosers. F. V. Smith (M '38) Sargent and Lundy, Engineers, Chicago, Ill. presided over the conference.

"Simplicity in Transformer Protection" by E. T. B. Gross (M '40) Illinois Institute of Technology, Chicago, Ill. This is essentially a review of the development and application of gas-operated relays as extensively used in Europe and in Great Britain for the protection of reactors and power transformers, an application for which differentially connected electromagnetic relays commonly are used in the United States. Whenever a fault in a transformer is beginning to form, heat is produced locally which leads to decomposition of insulating material with consequent production of gas. The nonelectrical relay is designed to be operated by the change in pressure within a transformer tank that is caused by the gas generated. Advantages of simplicity and sensitivity are claimed for the gas-operated relay.

"Rural Electrification in Iowa" by C. M. Stanley (M '40) Stanley Engineering Company, Muscatine, Iowa. In the nine years from June 30, 1936, to June 30, 1945, the rural consumers served in Iowa increased from 24,700 to 124,900. This electrical service is being provided primarily by privately owned utilities and by co-operatives, with a comparatively insignificant part by municipalities. Statistically, 39.8 per cent of the consumers are served by private utilities, 55.8 per cent by co-operatives, 3.2 per cent by municipalities, and 1.2 per cent by miscellaneous sources. Ten years ago an average monthly consumption of 100 kilowatt-hours per consumer was a goal toward which to work, whereas now the planning is in terms of from 300 to 700 kilowatt-hours per month per customer. An intense demand for electrical service by "practically every Iowa farm" is reported, including farmers who scornfully declined electrical service initially.

"Experience With Single-Pole Relaying and Reclosing on a Modern 132-Kv System" by J. J. Trainor (nonmember) and C. E. Parks (M '45) Public Service Company of Indiana, Inc., Indianapolis, Ind. In March 1941, the Public Service Company of Indiana placed in service, on an important 50-mile 132-kv intersystem tie line, a revolutionary installation of single-phase, or so-called single-pole, reclosing. This type of reclosing differs from the traditional three-pole reclosing in that, in the event of a single-phase-to-ground fault on the line, only the faulted phase is tripped open and then reclosed simultaneously at both terminals of the line. Experience has shown that for the short interval of time during which the faulted phase is open, either the parallel transmission line paths, or the ground itself, will assume the load current of the interrupted phase. Single-pole switching has permitted the transfer of large blocks of power over the lines so switched with less chance of system instability during faults which are confined to only one phase and ground, a condition which accounts for some 75 per cent of this company's transmission line faults for the type of line involved.

"Calculation of Power System Losses by Formula Using Constants Determined by the Network Analyzer" by J. B. Ward (A '40) Purdue University, Lafayette, Ind.

It is often necessary or desirable to determine the losses in a power system network for the purpose of forming operating schedules, for system planning, for interchange billing, or for other reasons. Ordinarily, the calculation of losses in a complicated power network is a lengthy and arduous task, perhaps involving a network-analyzer study of various operating conditions to determine the power flow and subsequently the making of line-by-line calculation of the I^2R losses. This paper presents a method whereby a formula for a given system is derived from a special network analyzer test, and the formula subsequently used to determine directly the total transmission loss for a large variety of operating conditions without further use of the analyzer. In this loss formula, the principle of superposition is utilized to let one set of analyzer studies serve to establish the line flow for a wide range of generating schedules with a given set of loads. The formula expresses the total loss as a function of station and tie-line kilowatt and kilovar loadings and corresponding bus voltages, information which is obtainable readily from station log sheets. The accuracy of this method is difficult to evaluate in general terms, inasmuch as errors depend to a large extent upon the assumptions and simplifications made in any specific case.

"Trends in Rural-Line Sectionalizing" by R. F. Quinn, (M '44) of the General Electric Company, Schenectady, N. Y. The enormous growth of farm use of electricity carries with it not only the responsibility but the economic necessity of assuring to farms a high quality of electric service. An electrified farm is an industry and it must be able to rely upon dependable electrical service. Ideal protection of rural lines would be obtained by the proper combination of oil circuit reclosers and fuse cutouts. The first step would be the application of enough oil circuit reclosers to handle all nonpersistent faults occurring anywhere on the system. Because the protective orbit of a recloser normally covers an extensive part of the system, this application of reclosers would result in relatively long sections of main lines, and would leave most branch-line taps unsectionalized. The second step, therefore, would be the application of enough fuse cutouts to supplement the reclosers by dividing the system into relatively short sections and by protecting all branch taps, so that a persistent fault would be isolated and its effects limited to a decidedly smaller section of the system that would be possible by reclosers alone. Proper co-ordination between the reclosers and the fuses would be required to assure that: (1) the reclosers handle all nonpersistent faults anywhere on the feeder; (2) the fuses do not blow on nonpersistent faults; (3) the fuses do blow to isolate persistent faults. Dependable co-ordination of this kind has been obtained on urban feeders, using station-type circuit breakers controlled by induction type vary-inverse-time-overcurrent relays. What the industry needs today is a recloser for rural line applications that will co-ordinate with fuses with the reliability of performance of station-type equipment.

CONFERENCE ON INDUSTRIAL POWER APPLICATIONS

John M. Webb (A '35) head of the department of electrical engineering, Eli Lilly and Company, Indianapolis, Ind., presided over the conference on industrial power applications. In general it could be concluded that electronic control of motor drives in conjunction with individual machine drives, is going to be more and more important to the industrial electrical engineer. The protection of plant equipment and distribution systems from lightning, and means of providing a flexible power distribution system also will receive considerable attention.

"Multiple Generator for Paper Machines" by G. E. Plaisted (A '46) General Electric Company, Schenectady, N. Y. It definitely is recognized that paper and paper products play a large part in our economics today. The usage is a bench mark to determine the level of our standard of living. The paper and paper products industry provides almost 400 pounds of this material annually per capita in the United States. This is several times the amount available in any other country. The potentialities of paper and paper products began to be recognized shortly after World War I. To increase production required more machines or machines of higher speeds or greater width. Chemical engineers developed processes for preparing the cellulose fibers in greater quantities and from materials heretofore considered inferior, and power engineers were faced with the problem of providing means of driving wider machines at higher speeds. For economic reasons, a paper machine was required to operate over wide speed ranges in order to make various weights and grades of paper. Mechanical drives involved line shafts, combinations of cone pulleys and belts for speed adjustment of individual paper machine sections, and gears and clutches for obtaining exact roll speeds and providing means of starting and stopping individual sections. Electrical engineers recognized the possibility of driving individual sections by individual motors, thereby eliminating heavy mechanical parts. Because its speed is a function of field flux and applied voltage, the d-c motor had inherent characteristics which made it particularly adaptable to a co-ordinated drive. The designers problem became one of maintaining an accurate speed relation among each of the several motors involved. The early sectional drives utilized one generator with individual motors, their speeds being maintained by field control. A further improvement in performance can be obtained by using an individual generator for each motor.

"Lightning Protection for Industrial Plants" by A. M. Opsahl (M '41) Westinghouse Electric Corporation, East Pittsburgh, Pa. On the basis of statistical averages, an industrial structure from 60 to 100 feet high and 100 by 100 feet in ground plan standing above a level plane may expect to receive a lightning discharge once in four or five years, or, in other words, one in some five such structures may expect a discharge each year. Surrounding objects will reduce the probability of direct hits, and the structural damage will depend upon the nature of the structure and protective installation if any. The installation of electric equipment used in industrial service usually is of the dry type with low factors of strength over normal voltage ratings. This fact makes it especially desirable with reference to industrial installations to use lightning rods and diverters to carry off direct strokes and surges, and to use lightning arresters specially designed and adapted to industrial service.

"Automatic Contouring Control" by J. M. Morgan, Jr. (nonmember) General Electric Company, Schenectady, N. Y. When used with the proper type of machine tool, this special type of control makes possible the machine duplication of work pieces of odd shapes by duplicating or reproducing the contour or shape of a master pattern or template. The electronic contouring control system described at Indianapolis combines basic electronic and magnetic circuits into a wholly electric and highly accurate system which provides constant speed of tool travel, great flexibility of control of machines, and continuous instead of step-by-step control. The electronic contouring control system operates as a "closed loop" system or, more specifically, as a "positioning follow-up" control system. Signal voltages generated in the tracing head by deflection of the tracing head stylus are used to operate the system. These signal voltages are fed into the contouring control desk where they are amplified and operated upon. The two resulting direct output voltages are fed into the motor-control circuits which in turn control the corresponding feed motors, which drive the tracing head—as well as the tool—so that the stylus follows the edge of the template. As there is a fixed mechanical relationship between the tool and the tracing head, and also between the work-piece and the template, the tool will duplicate on the work piece the contour of the template.

"A Demonstration of Industrial Distribution Systems" was given by H. G. Barnett (M '41) Westinghouse Electric Corporation, East Pittsburgh, Pa., displaying the comparative characteristics of radio and network distribution systems for industrial plants. The demonstration showed graphically, through the operation of a simulated industrial-plant system arranged on a panel board, that although the radial type of distribution system is less expensive to install and somewhat simpler physically, the network type of plant distribution system introduces an element of flexibility and freedom from interruption shutdown that very well may more than justify the initial instrument of expense. With the network system, any faults occurring in plant units are isolated quickly without interrupting other units; faults occurring between plant units are isolated without interrupting any units; repair, maintenance, and testing of elements of the plant distribution system can be accomplished at any time—even a transformer can be replaced—without interrupting the operation schedule of the plant.

CONFERENCE ON COMMUNICATIONS

The communication session was devoted to radio communication except for one paper on networks containing vacuum tubes. J. J. Pilliod (F '34) American Telephone and Telegraph Company, New York, N. Y. was chairman of the communications section.

"Manufacturing Problems of Air-borne Transmitter Equipment" by J. A. Green (A '42) Collins Radio Company, Cedar Rapids, Iowa. The various problems stemming from the peculiarities of aircraft installation lead to certain manufacturing problems. Factors of vibration, altitude which affects temperature and pressure, and power supply all must be taken into consideration.

"Phenomenon of Microwaves," a visual demonstration by H. R. Gruelle (nonmember) Indiana Bell Telephone Company, Indianapolis, Ind., supplemented the paper on the microwave relay system by showing some of the characteristics peculiar to microwaves.

"Analysis of Four-Terminal Networks Containing Vacuum Tubes" by Wilton R. Abbott (A '40) Iowa State College, Ames. Analysis and synthesis of four-terminal networks and systems of networks when these networks are linear, passive, and bilateral has been greatly facilitated by their study as a class without regard for their actual internal structure. It would seem that such a treatment of networks containing vacuum tubes would be of equal value. The method consists of determining four independent parameters of the circuit. Choice of equations depends on circumstances. In some cases if the complete circuit can be broken up into several interconnected four-terminal networks, it undoubtedly will be easier to determine the characteristics of each element and combine them.

"Mobile Radiotelephone Service" by J. G. Harden (nonmember) Indiana Bell Telephone Company, Indianapolis, Ind. Many of the requirements involved in providing the radio portions of mobile radiotelephone systems for use in common carrier service are similar to those encountered in city and state police and power company systems. Additional problems are imposed on a system which must provide satisfactory communication between mobile units and instruments in a regular wire network. Most of the special requirements of such a system are incorporated in a control terminal which provides for combination of two radio-derived one-way transmission paths into a single two-way circuit; combination of various receiver outputs delivered to the terminal by wire lines; introduction of fixed gain as required by transmitting and receiving branches; introduction of automatic gain in transmitting branches to maintain optimum modulation; provision of necessary features for switchboard operation; monitoring of emissions and frequency of the stationary transmitter; and provision of various test functions. These requirements are those of a primarily urban system.

"A Multichannel Microwave Radio Relay System" by H. S. Black (F '41), J. W. Beyer (M '36), T. J. Grieser (nonmember), F. A. Polkinghorn (M '39) Bell Telephone Laboratories, Inc., New York, N. Y. An eight-channel microwave relay system used by the Army and Navy uses frequencies approaching 5,000 megacycles. The waves are concentrated into a sharp beam and do not travel along the earth much beyond seeing distances. Because the waves are sharply focused other systems in the vicinity can operate on the same frequency and only one four-millionth of the power required by a nondirectional antenna is needed. Although this equipment was designed for military use its basic principles and design features can be applied to telephone and telegraph communication. Sharply beamed radiation and reception, pulse modulation, and time division multiplex are all factors which make the system attractive in competition with other communication systems. *Electrical Engineering* November 1946, volume 65, page 516, carries an article on this subject by F. B. Bramhall (F '44) The Western Union Telegraph Company, New York, N. Y. on "Radio Relays for Telegraphy."

CONFERENCE ON INDUSTRIAL CONTROL

Two papers which evoked especial interest at this meeting were the ones on induction motor control and a-c magnets. The paper on induction motor control was apparently based on much development work and represents one approach to the problem of variable speed a-c motor drives. Possibilities which drives of that general nature offer undoubtedly will lead to further development involving induction motors under unbalanced conditions. The paper on a-c magnets presented much information with regard to a-c magnet performance which is not generally known.

Session Chairman R. W. Jones (A '42) associate professor of electrical engineering, Northwestern University, Evanston, Ill, stated that the gap which exists between textbooks and periodical literature is a difficult hurdle for the young engineer and that it might be advisable for the Institute to sponsor sessions of an educational nature on a variety of specialized topics at the annual meetings. He also observed that there is little material available on a-c magnet design outside the field and that the Institute could well serve the members if publication of such material could be arranged.

"Trends in Co-ordinated Control" by G. A. Moffett (M '44) General Electric Company, Schenectady, N. Y. Ranking very highly in the long list of control activities are "co-ordinated control equipments" which have offered great improvements through increased output in continuous processes such as for paper, textiles, rubber, and steel. With mechanical systems comprising single motors, line shafts, gear reducers, couplings, and power take-offs, the devices for adjusting speed, entry, wind-up, tension, and so forth are not self-regulating and frequently must be readjusted manually. The electronic sectionalized drive has replaced the mechanically coupled system and makes possible a more simplified and flexible setup which is self-regulating and continuously adjustable. Co-ordinated control equipment has many applications other than sectionalized drive. These include large wind tunnels with thousands of horsepower on a single shaft, model testing, power absorption and recovery systems, chemical and petroleum processes, printing presses, and automatic furnaces. During the war a large number of control systems were devised to accomplish certain functions. The experience gained has enabled the best systems to be chosen for specific applications. The next move is one of standardization of functions and their circuit components without hindrance to flexibility of application.

"Control of Slip Ring Motors by Means of an Unbalanced Primary Voltage" by N. L. Schmitz (A '43) Cutler-Hammer, Inc., Milwaukee, Wis. Present applications of the unbalanced primary speed control for slip ring induction motors include crane and drawbench installations where a high lowering or return speed and braking action for slowdown is desired. To obtain the required unbalanced voltages for operation an autotransformer suitably tapped is used. As is true of any apparatus, this equipment has limitations. Most talked of limitation is heating of the motor. Current values of 200 to 250 per cent are encountered in the primary phases at slow speeds, but this is also true of armature loop currents in d-c motors. Even with a more severe duty cycle than the 15 seconds on and 45 seconds off specified by the American Standards Association undue heating is improbable, because coils adjacent to the hot winding are cooler thus facilitating heat transfer, and temperature equalizes rapidly when the motor is operating with balanced currents under load or when it is at rest. This is claimed to have proved true in tests as well as in theory. A wider speed range is possible if capacitors are used in the secondary circuit in place of the usual resistors. Use of inductance in series with the capacitance in the secondary permits further reduction in speed. In order to obtain large secondary current at high power factor for maximum torque at low speed the values of inductance and capacitance are selected to approach series resonance at the frequency of induced secondary currents.

"Problems in Design of A-C Magnets and Solenoids" by L. T. Rader (M '43) Illinois Institute of Technology, Chicago. Although a-c magnets have existed for some time, little has been published about their design. An attempt is made to present for consideration some of the difficult problems involved. To facilitate a strictly analytical solution for the design of a-c magnets several assumptions are usually made: flux varies directly with current, saturation is disregarded; leakage flux is neglected; iron losses are neglected; voltage drop is neglected. These assumptions are so bad that a solution dependent on them is very misleading. In order to design a magnet certain specifications must be furnished the designer. He must know the load displacement curve, duty cycle, quiet pull required in the sealed position, mechanical life, dimensions to which the device must conform, and the maximum volt-amperes which can be drawn from the line. The design problem consists of two parts—conditions when the magnet is closed or sealed and conditions when the magnet is open. The problems encountered with the magnet in the closed position are all involved with those factors producing heat. Consequently voltage and steel are important in these considerations. For the open position force of the magnet is a major consideration and the leakage path assumes great importance. Few data are available on factors affecting leakage and volt-ampere data are used as a substitute. A rigid requirement of magnets in the closed position is that they must be quiet. Saturation is involved in this problem and it immediately deviates from a rigid analytical solution. Also important is the mechanical life. Rapid acceleration which results from the device being able to operate at 85 per cent voltage produces high impact forces. Repeated application of such forces introduces many problems of mechanical nature to design.

CONFERENCE ON BASIC SCIENCES

A variety of subjects was covered in the session on basic sciences, of which K. W. Miller (M '29), Armour Research Foundation, Chicago, Ill., was chairman.

"The Electronic Numerical Integrator and Computer" by T. K. Sharpless (A '44) University of Pennsylvania, Philadelphia, Pa. Need of the United States Army Ordnance Department Ballistic Research Laboratories at Aberdeen, Md., for a machine capable of rapidly performing calculations for preparation of firing tables initiated work on the electronic numerical integrator and computer. The machine, built at the University of Pennsylvania, Moore School of Electrical Engineering, was made sufficiently flexible to enable it to be applied to a variety of problems other than exterior ballistics, which involve numerical integration. Considerable work is required to prepare problems for solution but once the mathematicians have done this the machine performs the computations extremely rapidly. Comprising the machine are 20 accumulators, which perform the functions of addition, subtraction, and storage of numbers; 3 function tables, which permit looking up of functional values of empirical data; a high-speed multiplier; a divider-square rooter; a constant transmitter, which works with an International Business Machines card reader; and a transmitter which uses an IBM card punch. Also provided are the master programer, cycling unit, and initiating unit which control the operations of the machine.

"Radioactive Isotopes and Their Applications" by G. Freidlander (nonmember) General Electric Company, Schenectady, N. Y. Radioactive tracers have been employed for several purposes which include study of the behavior of minute quantities of material; study of exchange reactions; tracing of an element through chemical changes, which is useful in organic syntheses, analytical separation processes, and a multitude of biological researches; and for following migration of material. The latter purpose has found use in diffusion studies, investigations of corrosion and wear, and applications to process control. Limitations in application are determined by the half-life of the isotope, the chemical form and amount in which it is available, the type and energy of radiation, and the activity per unit weight necessary. The method by which an isotope is produced must be considered because it usually determines the available specific activity.

"Latest Developments in High Energy Physics" by Karl Lark-Horovitz (nonmember) Purdue University, Lafayette, Ind. As an outcome of war developments many physicists have become interested in electronic devices, and particularly a great many physicists whose main interest was and is in the field of nuclear and high energy physics or cosmic rays. These men, using techniques developed during the war, have shown that new devices can be built using such engineering methods as have been available in the high-frequency fields in radar development. Applying these ideas successfully to the acceleration of particles, they designed and built devices which promise to produce high energy protons of the order of 40 to 100 million volts in the so-called linear accelerator, high energy electrons in a device which is called the synchrotron, and high energy deuterons in the frequency-modulated cyclotron. These developments promise to open an entirely new field in nuclear physics research.

"Solar Electron Radiation and Its Effect on Power Transmission and Communications" by Doctor J. T. Wilson, Allis-Chalmers Manufacturing Company, Milwaukee Wis. Almost everyone has witnessed the effects of sunspot activity in disruption of radiobroadcasting during magnetic storms. The effect of solar electron radiation has had its effect on power transmission as well.

National Electronics Conference Papers Digested

For the purpose of giving AIEE members a brief cross section indicating the scope and nature of the 57 scheduled technical subjects discussed at the second National Electronics Conference in Chicago, October 3–9, 1946, the following brief digests are presented, drawn primarily from information furnished by the authors. Except to the extent that individual authors may have extra copies of their papers these papers are not individually available. All these papers are scheduled to be published in the "Proceedings" of the second National Electronics Conference, which currently is in production and is expected to be available for distribution early in 1947. Persons interested in securing copies of the "Proceedings" should take action immediately, forwarding a request *and* a remittance of $3.50 to E. H. Schulz, secretary, National Electronics Conference, Inc., Technology Center, Chicago 16, Ill.

TELEVISION

"Color Television—Latest State of the Art," by P. C. Goldmark (M '45) Columbia Broadcasting System, was presented as an oral dissertation covering calculations and performance characteristics of the equipment and procedures currently involved in developing color television.

"Westinghouse Color Television Studio Equipment" by D. L. Balthis of the Westinghouse Electric Corporation, described equipment patterned after the color television studio equipment designed and developed by the Columbia Broadcasting System in New York, N. Y. The studio equipment consists of the electrical and optical equipment required to convert a 35 millimeter color slide, or a 16 millimeter color film and its associated sound, into signals suitable for input to an ultrahigh-frequency color television transmitter utilizing three primary colors. Sound and picture signals are transmitted on the same carrier frequency, with special arrangements for color synchronization.

"Television Transmitter for Black-and-White and Color Television" by Norman Young of the Federal Telecommunication Laboratories was descriptive of the equipment indicated.

"Stratovision System of Communication" by C. E. Nobles of the Westinghouse Electric Corporation and W. K. Ebel of The Glenn L. Martin Company was descriptive of the experiments and plans that are being directed toward the development of a series of air-borne television transmitting stations intended to serve most of the United States from a small number of points. It is contemplated that a network of eight airplanes flying at a height of 30,000 feet and some 400 miles apart will be able to perform a service that would require 100 ground relay stations 35 miles apart.

"The Electrostatic Image Dissector" by H. Salinger of the Farnsworth Television and Radio Corporation describes a tube which uses magnetic focusing but which achieves scanning by electrostatic rather than magnetic deflection. The theory of this tube as worked out indicates that distortions are greatly reduced as compared with the conventional method of operating a dissector. The scanning generators are of the so-called "bootstrap" type. Deflection is accomplished by means of several wires (12 in the present tube) disposed axially inside the tube. The two scanning sawtooth voltages are applied in different proportion to each of these wires.

"The Use of Powdered Iron in Television Deflecting Circuits" by A. W. Friend (M '39) of the Radio Corporation of America. The horizontal deflection of electron beams in television systems heretofore has required excessive energy dissipation and expensive circuit components. For deflection by magnetic means, the deflection transformer and yoke have presented serious problems in the economical design of television receivers. Now, low-loss systems have been constructed which require no additional electric energy and which provide large increases in deflection capability. Also, the development of transformer and yoke cores molded from powdered iron have resulted in appreciable cost reductions as compared with sheet or strip-metal types. Molded core structures produce negligible acoustic radiation in comparison with laminated core structures. A low-cost system has been constructed to provide deflection of a 27-kv electron beam, yielding maximum picture size on a 60-degree kinescope driven by a single 6BG6G beam tetrode.

"Television Equipment for Guided Missiles" by C. J. Marshall, Wright Field, Ohio, and Leonhard Katz of the Raytheon Manufacturing Company, outlined experiments with lightweight transmitting equipment begun as early as 1941 by the Army Air Forces at Wright Field.

ELECTRONIC INSTRUMENTATION

"A Method for Changing the Frequency of a Complex Wave" by E. L. Kent of C. G. Conn, Ltd. This method is applicable in tone analysis, tone synthesis, psychological studies in tone quality, and other similar studies. The frequency of a complex wave may be changed by the use of a special cathode-ray tube or by the use or ordinary radio tubes utilized in special circuit arrangements.

"Detectors for Buried Metallic Bodies" by L. F. Curtis (F '29) of the Hazeltine Electronics Corporation, briefly outlines some of the problems encountered in the development of a detector for the Engineer Corps of the United States Army under the supervision of the National Defense Research Council.

"The Pressuregraph" by A. Crossley of the Electro Products Laboratories, Inc. This device is an electronic instrument for indicating in a linear manner the static or dynamic pressure of an internal combustion engine, pump, or other pressure device. The Pressuregraph consists of a diaphragm upon which the pressure waves impinge, the diaphragm being part of an electric capacitor. Movement of the diaphragm results in change in capacitance which in turn actuates a bridge or electronics circuit causing modulation of a high-frequency carrier, the modulated wave being amplified and delivered to a cathode-ray oscillograph which in turn presents a visual translation of the pressure-time picture that is the indicator card of the device being investigated.

"The Notch Wattmeter for Low-Level Power Measurement of Microwave Pulses" by D. F. Bowman of the Hazeltine Electronics Corporation presents a review of the basic methods of microwave power measurement. In the "notch wattmeter," a method is employed whereby the radio-frequency pulse of unknown amplitude is matched in amplitude with an interrupted continuous-wave signal. The interrupted portion, or notch, of the continuous-wave signal is adjusted to equal in length and to coincide in time with the unknown pulse. This adjustment facilitates accurate matching of amplitudes. The amplitude of the combined signal, as measured by an average power-indicating instrument, therefore is an accurate measurement of the peak pulse power of the unknown radio-frequency signal.

"The Mechanical Transients Analyzer," by G. D. McCann (M '44) of the Westinghouse Electric Corporation presented a brief description of the Westinghouse Electric analog computer which was developed for the solution of complex algebraic and and differential equations. The differential equations specifying the physical system are represented by electric circuit parameters such as resistors, inductors, and capacitors. The variables of the system are represented by currents and voltages. Electronic amplifier circuits are used to represent negative impedance elements or energy sources, such as required for servomechanisms. Special electronic circuits and techniques have been developed to produce arbitrary steps of excitation functions and initial or boundary conditions.

"High-Performance Demodulators for Servomechanisms" by K. E. Schreiner of the Servomechanism Laboratory, Massachusetts Institute of Technology. As the frequency-response range of servomechanisms is extended, the necessity for minimizing time delays in the response of individual components correspondingly is increased. In many instances, in servomechanisms using a-c data transmission, one of the important time delays is that introduced by the demodulator—or phase-sensitive detector—which in addition to small time delay should have the additional characteristic of low ripple component of output voltage. Normally, either of these characteristics may be attained at the expense of the other. A particular high-performance circuit is described and analyzed, the principal feature of which is the combination of small time delay and low ripple components of output voltage. The detecting element is a double-triode, the grids of which are excited in such a way as to establish a bidirectional conduction path between input and output for a very short period of time while the signal voltage is passing through its peak. The unique feature of this type of demodulator is essentially that the output voltage is not allowed to change except at the instant of peak signal input.

"The Theory and Design of Several Types of Wave Selectors" by N. I. Korman of the Radio Corporation of America. A wave selector is a device for sampling, measuring, or injecting a traveling wave in a transmission line or wave guide. As a knowledge of the traveling waves on a transmission line is equivalent to a knowledge of the standing waves, and as the wave selector can measure the traveling waves without the use of any mechanically moving parts, it can supplement and in many cases replace the more complicated standing wave detector. The principles of operation of the wave detectors known as the long-slot, electromagnetic, reversed-coupling, and phase-indicating types are given, and applications discussed.

"An Oscillographic Method of Presenting Impedances on the Reflection Coefficient Plane" was described in some detail by A. L. Samuel (M '34) of the University of Illinois.

"Electron Optics of Deflection Fields" were described by R. G. E. Hutter of Sylvania Electric Products, Inc. The properties of electric and magnetic deflection fields commonly used in cathode-ray and television tubes are discussed and mathematical theory expressions for the field quantities are given. If the assumption be made that the deflection of electrons from the optical axis is small while the electrons are under the influence of the deflecting field, the problem of finding general mathematical expressions for the deflection and for the deflection defects can be solved by two methods—the "path method" and the "Ikonal-method"—which are known from the theory of electron lenses. The resulting general expressions may be applied to a number of special fields.

Trends in cathode-ray oscilloscope design were discussed by W. L. Gaines of the Bell Telephone Laboratories, Inc. War demands for the development and servicing of radar and other equipment, and future demands now foreseeable, require signal channels which include attenuators, delay networks, and amplifiers capable of handling transient wave forms, and precision slave start–stop sweeps with a high ratio of operate–restore time. Accessory circuits such as lockout, sweep delay, and timing circuits have been found to be extremely convenient for many purposes and essential for some purposes. Inherent difficulties include adequate delay networks and signal attenuators for the signal channel, and adequate unblanking for the sweep circuit. Also, cross talk between circuits within the oscilloscope requires careful attention to grounding paths and to bypassing as well as shielding.

DIELECTRIC AND INDUCTION HEATING

"Microwaves and Their Possible Use in High-Frequency Heating" by T. P. Kinn

(M '44) and J. Marcum discusses the sources of power now available in the microwave region (1,500 to 30,000 megacycles) giving available power output and other pertinent characteristics. Coaxial transmission lines and wave guides were explained in simple terms.

"Ignitron Converters for Induction Heating" by R. J. Ballard and J. L. Boyer (A '43) of the Westinghouse Electric Corporation described a new type of ignitron frequency converter for melting-furnace and forging-heater applications. Derived from the d-c to a-c parallel inverter circuit, which has been known and used for many years, this new circuit is called the "cyclo-inverter." Three-phase 60-cycle power is converted to single-phase power at a higher frequency by means of a single conversion.

"Dielectric Preheating in the Plastics Industry" by D. E. Watts, G. F. Leland (M '45), and T. N. Willcox (A '38) of the General Electric Company. Dielectric preheating is a comparatively new industrial tool in the plastic industry. By this means, heat is generated within plastics materials just prior to molding, by applying a high-frequency electric field. The "preform," thus being uniformly soft, allows new techniques to be used in the processing of thermosetting plastic materials.

"The Problem of Constant Frequency in Industrial High-Frequency Generators" by E. Mittelmann of the Illinois Tool Works is related to the recent frequency allocation for medical and industrial generators by the Federal Communications Commission which emphasized the problem of constant-frequency generators. The requirement for constant frequency poses two problems:

1. To maintain a constant frequency operation.
2. To force the generator to this frequency.

A solution has been found in the combination of a sharp-cutoff high-pass filter and a sharp-cutoff low-pass filter providing control voltage outputs of different polarities for frequency deviations above and below the assigned frequency.

INDUSTRIAL APPLICATIONS

"Large Electronic D-C Motor Drives" by M. M. Morack (M '42) of the General Electric Company describes the application of electronic control equipment to d-c motors of 600 horsepower. The electronic drive system provides control means including both the starting and the running of the drive motor. The functions are equivalent to the Ward Leonard system, in that an ignitron rectifier replaces armature voltage adjustments by field control.

"Electronic Speed Control of A-C Motors" by W. H. Elliott of Cutler-Hammer, Inc., reviews briefly some of the basic electronic circuits which may be employed to control the more common types of a-c motors, such as the drag-cup shaded-pole, universal, and wound-rotor motors.

"The Electronic Method of Contouring Control" by J. Morgan of the General Electric Company. Contouring control is basically the following of a master template or pattern by a stylus or tracer. Using signals generated by the tracer, other portions of the contouring system control the mechanism, so that the cutting tool reproduces the contour of the template on the work. This electronic system uses a new combination of basic electronic circuits to obtain the desired controls.

"Production Test Facilities for High Power Tubes" by W. L. Lyndon and B. Sheren of the Radio Corporation of America described an equipment and installation which was engineered and designed especially for the testing of power tubes. The physical size of high power electron tubes, plus the magnitude of voltages, currents, and power outputs encountered, necessitate the design of special equipment for the testing of these tubes. Flexibility in such equipment is essential.

ANTENNAS AND WAVE PROPAGATION

"Radio Propagation at Frequencies Above 30 Megacycles" by K. Bullington of the Bell Telephone Laboratories, Inc. Radio propagation is affected by many factors, including the frequency, distance, antenna height, curvature of the earth, atmospheric condition, and the presence of hills and buildings. The influence of these factors is represented in nomograms, by means of which an estimate of the received power and the received field intensity for a given point-to-point radio transmission path ordinarily can be obtained in one minute or less.

"Interference Between Very High Frequency Radio Communication Circuits" by W. R. Young, Jr., of the Bell Telephone Laboratories, Inc. Interference between different radio circuits is an old problem, one which generally has been solved in the past by trial and error and by "hand-tailored" special filters and other equipment. With the general increase in the usage of radio communication, the amount of potential interference is greatly increased. Generally there is a large difference between transmitting and receiving energy levels. As a result, spurious radiations, spurious responses, and lack of sufficient receiver selectivity may in many instances cause interferences. Common causes of such interferences are discussed. Sample measurements are given to illustrate the relative magnitude of the various modes of behavior. Formulas are given which permit computation of the frequency of these disturbances. A method is described for making charts suitable for a given type of equipment from which the spurious frequencies can be read directly as a function of the operating frequency.

"Aircraft Antenna Pattern Measuring System" by Otto Schmitt, Airborne Instrument Laboratories. This is a "packaged system" for measuring the radiation patterns of aircraft antenna by the use of scale models.

"Improvements in 75-Megacycle Aircraft Marker Systems" by B. Montgomery of the Northwest Airlines describes improved aircraft marker systems which permit air line pilots to follow their courses to their destinations with greater precision. The system consists of transmitters operating on 75 megacycles with antenna systems of a type which concentrate the radiation in an upward direction, the markers thus indicating the airplane position in relation to range stations on the ground.

"Problems in Wide-Band Antenna Design" were discussed by A. G. Kandoian (A '40) of the Federal Telecommunication Laboratories. With the increased use of higher and higher frequencies, the problem of wide frequency range operation of antennas of various types assumes more importance, and its solution becomes more feasible. The most general requirement is that both the impedance and radiation patterns be essentially independent of frequency over the operating range, whether applying to nondirective antennas or to highly directive antennas.

"Slot Radiators" were described and discussed by A. Alford (M '42) consulting engineer, Chicago, Ill.

"Results of Field Tests on Ultrahigh-Frequency (490 Megacycles) Color Television Transmission in the New York Metropolitan Area" were reported by W. B. Lodge of the Columbia Broadcasting System. Since January 1946, the CBS has operated an ultrahigh-frequency color television transmitter located on top of the Chrysler Building in New York City. Surveys utilizing mobile recording field intensity equipment and involving operation of actual television receivers at a large number of different locations throughout the metropolitan area, carried on continuously since the transmitter began operation, have provided considerable qualitative and quantitative data on:

1. The extent of the service area.
2. The effect of changes in height of a receiving antenna.
3. Shadowing.
4. Effect of trees and foliage.
5. Stability of transmission.
6. Location and extent of areas receiving multipath transmission.
7. Efficacy of methods of eliminating "ghosts" in areas receiving multipath signals.

A method of predicting possible areas of multipath transmission has been developed.

FREQUENCY MODULATION

This was a panel discussion on the design of frequency modulation receivers.

"A Permeability-Tuned 100-Megacycle Amplifier of Specialized Coil Design" was described by Z. Benin of Zenith Radio Corporation. Small coils were designed that could be positioned between tube and switch circuit components to make a compact radio-frequency tuner with a minimum of stray inductances. Powdered-iron-core permeability tuning improves the frequency drift with temperature. A variable-pitched winding technique was used to straighten out the tuning curve of the core-and-coil combination. The lengths of core-and-coil, the winding, and the powdered-iron materials were designed to give a logarithmic frequency travel curve. By mathematical analysis such a characteristic was shown to enable the elimination of trimmer capacity circuit adjustment and the substitution of an initial setting of the tuning core. The result was a space-saving and frequency-drift improvement.

"Very High Frequency Tuner Design" by G. Wallin and C. W. Dymond of the Galvin Manufacturing Company describes the design of a tuner for use in the 88–108-megacycle frequency modulation band. The circuit used is essentially an adaptation of a single-oscillator double superheterodyne. One section of a dual triode tube performs the function of oscillator and first converter. The other section is used as the second converter. Three very "Hi-Q" elements used for tuning the antenna, oscillator, and variable intermediate frequency, are provided by means of permeability tuned transmission lines. This adaptation of the transmission line permits the use of an effective 250 micromicrofarad oscillator tuning capacitance, with attendant reduction in warm-up drift. A high order of temperature stability also is achieved.

"Front-End Design of Frequency Modulation Receivers" by C. R. Miner of the General Electric Company describes the methods used by the General Electric Company in solving the practical problems involved in the design of the front-end of a frequency modulation receiver, involving the use of a new type of variable inductance tuner, together with a unique mechanical design.

"A Single-Stage Frequency Modulation Detector" by W. E. Bradley of Philco Radio and Television Corporation describes the equipment and circuit combinations developed by that company.

"Frequency Modulation of High-Frequency Power Oscillators" by W. R. Rambo of the Airborne Instrument Laboratories. Practical cases arise in which it is desirable to frequency-modulate power oscillators operating at high frequencies, and thereby to avoid the use of amplifying and frequency-multiplying stages. A common-grid reactance-tube circuit is described that has an inherent tendency to reduce incidental amplitude modulation, permits the use of triode reactance tubes in the very high frequency and the ultrahigh frequency ranges, and has a configuration that permits its incorporation in ultrahigh frequency common-grid coaxial-line oscillator circuits. Lighthouse tube oscillators have been modulated over a plus or minus one-half per cent band, at 1,000 megacycles by this means; much wider band widths are possible in the very high frequency range.

AIR NAVIGATION SYSTEMS

"Automatic Radio Flight Control" by F. L. Moseley, Collins Radio Company, and C. B. Watts of the Federal Telecommunication Laboratories discusses the general problem of automatic control of aircraft flights on radio-defined tracks as an aid to point-to-point navigation, traffic control, and final approach and landing. A brief outline is given of the various radio navigational facilities which are available to define suitable tracks for automatic aircraft guidance, and the paper describes various systems which have been tested extensively. Widespread adoption of automatic radio flight control systems is recommended as an indispensable aid to all-weather operation of aircraft.

"Naviglobe—Long-Range Air Navigation System" by P. R. Adams and R. I. Colin, Federal Telecommunication Laboratories describes a directional transmitting type of system developed and proposed by the Federal Telecommunication Laboratories for the particular requirement of long-distance transoceanic air navigation. By virtue of its particular antenna disposition, and by using all quantitative signal relationships and not merely the conventional equisignal indications, the coverage is continuous and omnidirectional. Line-of-position indications are automatic and visual, corresponding to the true bearing of the observer, without the need of special charts. The fundamental features are designed on the basis that a minimum ground-station range of 1,500 miles is required to cover the North Atlantic area and the other likely transoceanic and polar routes. The study indicates that extremely reliable propagation at such ranges is to be expected most feasibly and economically if low frequencies in the neighborhood of from 70 to 100 kilocycles are used.

"Teleran—Air Navigation and Traffic Control by Means of Television and Radar" was presented by D. H. Ewing and R. W. K. Smith of the Radio Corporation of America. Wartime developments of radar technique offer a new approach to the problem of improving air navigation and traffic control, two fields in which existing equipment is obsolescent. Utilizing the special qualities of different special equipment developed originally for military service, the new proposed Teleran system presents aircraft position information to ground observers and controllers on a series of plan-position indicators. One indicator is used for each altitude layer, and is superimposed on a map of the region covered by the ground radar. This information, together with desired information of other factors, including weather and traffic control, is transmitted by television to each aircraft in the region. Each co-operating aircraft is equipped with a transponder beacon which serves not only to reinforce the radar echo, but also to provide an altitude-dependent reply which allows the ground station operators to differentiate among aircraft according to altitude.

RADIO RELAY SYSTEMS

"A Microwave Relay Communication System" by G. G. Gerlach of the Radio Corporation of America reviews the experimental results obtained with a 4,000-megacycle radio relay system connecting New York and Philadelphia. This system employs a frequency-modulated subcarrier which in turn is used to frequency-modulate the final carrier. Demodulation to the subcarrier frequency is effected at relay stations. Microwave relay equipment resulting from this experimental work which will be installed by the Western Union Telegraph Company in a circuit connecting New York, Washington, and Pittsburgh is described.

"Pulse-Time Multiplex Broadcasting of the Ultrahigh Frequencies" by D. D. Grieg (A '39) and A. G. Kandoian of the Federal Telecommunication Laboratories describes the result of operating experiences with a system in experimental operation in New York City. The system combines a total of eight high fidelity programs operating on the frequency of 930 megacycles; the pulse-time video band width utilization is 2.8 megacycles; base pulse repetition is 24 kilopulses per second per channel, with a time modulation deviation of plus or minus one microsecond.

"The Cyclophon—a Multipurpose Beam Switching Tube" by J. J. Glauber, D. D. Grieg, and S. Moskowitz (A '41) of the Federal Telecommunication Laboratories concerns the theory of operation of beam switching tubes in general, a type of tube originally developed for pulse-time modulated communication systems, but which proved to have many other interesting applications. The tube described employs secondary emission, and the paper discusses the characteristics of these dynodes and their equivalent circuits. Construction and processing of the tube are described in detail, and the results of life and vibration tests also are given.

MICROWAVE GENERATORS

"Continuous-Wave Ultrahigh-Frequency Power at the 50-Kw Level" by W. G. Dow (M '32) and H. W. Welch of the University of Michigan describes the theory and use of electronic apparatus capable of generating continuous-wave radio-frequency power at the 50-kw level in the frequency range from 350 to 600 megacycles. This equipment was developed originally during the war to meet certain military requirements for the jamming of German radar equipment.

"Microwave Frequency Stability" by A. E. Harrison of the Sperry Gyroscope Company. A review of the history of previous expansions of the radio-frequency spectrum emphasizes the importance of frequency stability in any region. The microwave region is no exception. Several methods are available for providing the frequency stability that is essential throughout the radio frequency spectrum, including the microwave region. Direct crystal control of the microwave power can be obtained by using klystron frequency multipliers and klystron power amplifiers. Automatic frequency control systems also may be used. The reference frequency may be either a crystal-controlled oscillator or a precision cavity resonator. All these methods are compared, and their advantages and disadvantages discussed.

"An All-Metal Tunable Squirrel-Cage Magnetron" by F. H. Crawford of Williams College describes the "donutron" which is tuned by the relative axial displacement of alternate anode segments through the flexure of one wall of the cavity in which the anode structure is supported. The best model to date tunes over a 1.5:1 ratio with power flat to 3 decibels. A single value of voltage and magnetic field is adequate for the entire tuning range. Of the various modes of operation, two are important: A long-wave tunable cavity mode, and a short-wave resonant re-entrance-line mode. The former can be suppressed entirely, and the latter can be enhanced by a special phase-reversing anode. In the line mode, the best powers have been observed. In cold tests, modes of 4 centimeters as well as indication of even shorter resonances have been found.

"Design of Wide-Range Coaxial Cavity Oscillators Using Reflex Klystron Tubes" by J. W. Kearney (A '44) of the Airborne Instrument Laboratories describes the techniques utilized in the design of these tubes. The tuning range of each oscillator is approximately two to one in frequency in the 1,000- to 11,000-megacycle ranges. A survey of the general requirements of such oscillators for use as local oscillators in superheterodyne receivers and as signal sources in test equipment is followed by a discussion of the suitability of reflex klystron tubes for those oscillators in the particular frequency range. The operation of reflex klystrons is described with regard to their use over large frequency ranges, and adaptation of present-day tubes to coaxial resonators is discussed. Design characteristics are outlined for optimum cavity dimensions, suitable contacts to the tubes, noncontact-short-circuiting tuning plungers, and output coupling devices.

THEORETICAL DEVELOPMENTS

"Bunching Conditions for Electron Beams With Space Charge" by L. Brillouin of the Cruft Laboratory, Harvard University. A rigorous computation of electron trajectories within plane structures may be made, with the inclusion of space charge effect, by use of the Llewellyn method of integration. These computations can be used for a discussion of conditions leading to intercrossing trajectories that is bunching. The discussion is carried out in this paper for a conventional plane diode, for a diode wherein electrons are injected with a given initial velocity, for a velocity-modulated beam, for a plane magnetron, and for a plane magnetron with velocity modulation.

"Generalized Boundary Condition in Electromagnetic Problems" by S. A. Schelkunoff (M '34) of the Bell Telephone Laboratories, Inc. The conception of an idealized presently conducting boundary long has been helpful in simplifying electromagnetic problems. The paper deals with a generalization of this concept and particularly with its application to the theory of the magnetron.

"Conformal Transformations in Orthogonal Reference Systems" by C. S. Roys (M '45) of the Illinois Institute of Technology. Some general equations for conformal transformations which correspond to the Cauchy-Riemann Equations for Cartesian co-ordinates. These then are applied to shielding and to electron tube problems involving recurrent structures, leading to an equivalent, physically realizable unit structure. This removes the necessity of obtaining an appropriate transformation by cutting and trying.

INFRARED AND MICROWAVE SYSTEMS

"Reflex Oscillators for Radar Systems" by J. O. McNalley and W. G. Shepherd of the Bell Telephone Laboratories discusses problems encountered in the design of a series of reflex oscillators for military application. As an essential element of a radar receiver, a beating oscillator is required to heterodyne the received signals to intermediate frequency. Military requirements dictated the necessity of compact simple devices for such an oscillator. The reflex oscillator, because of its single-cavity tuning control and vernier tuning provided by the electronics tuning, provided a very satisfactory solution to these problems.

"Modulation of Infrared Systems for Signaling Purposes" by W. S. Huxford, of Northwestern University. In general, the sources used in "light" beam communication systems are modulated in one of three ways. The first method—mechanical modulation—utilizes shutters or mechanically vibrated elements to secure variation in intensity of the radiated beam after the beam leaves its source. In the second or optical method of modulation, some form of optical shutter such as a Kerr cell is used. The third method consists of varying the radiated flux generated in the source by electrical modulation of the power supplied to the lamp. A detailed description is given of the mechanical and optical methods of modulation used in captured German equipment. As an example of electrically modulated sources, the characteristics of a new source called the "concentrated-arc" lamp are described.

"Photo Detectors for Ultraviolet, Visible, and Infrared Light" by R. J. Cashman of Northwestern University considers photoelectric cell developments of the past few years, including photoemissive, photovoltaic, and photoconductive types. Photoemissive cells with caesium alloy cathodes have great sensitivity in the ultraviolet and the visible regions of the spectrum. Other cells of this type with pure metal anodes have considerable application in the ultraviolet region. Present-day photovoltaic or self-generating cells of the selenium type, most commonly used for work in the visible spectrum, are characterized by high output and good stability. Two new photoconductive cells of thallous sulphide and lead sulphide, which recently have been released, exhibit marked infrared sensitivity and excellent stability.

"Military Applications of Infrared Viewers" by G. E. Brown of the Engineer Board, Fort Belvoir, Va., described several items of infrared equipment developed by the Engineer Board. These items included the "Metascope," which employs an infrared sensitive phosphor for detecting sources of infrared radiation, and an infrared telescope using the 1P25 image tube for viewing areas which are "illuminated" by infrared radiation.

SPECTROSCOPY AND MEDICAL APPLICATIONS

"The Use of Radioactive Materials in Clinical Diagnosis and Medical Therapy" by J. T. Wilson of the Allis-Chalmers Manufacturing Company, described the methods for the production of radioactive isotopes and reviewed methods and equipment currently in use for inducing artificial radioactivity by means of bombardment with neutrons, alpha particles, protons, and gamma radiation. He explained the close relationship existing between the growing knowledge of nuclear phenomena and medical research pertaining to both diagnosis and treatment of human ailments. The use of radiophosphorus and radioiodine in clinical diagnosis and biological tracer techniques was reviewed. By the use of known and predetermined dosages of radioactive matter, it is possible now to trace accurately and definitely the transition and assimilation of such materials and of the materials with which they are mixed by both animal and vegetable tissue. With the door to an accurate knowledge of atomic structure opened by the results of recent nuclear research, and with the opportunity thus presented for discovering and studying the similarities, differences, and interrelationships of the atomic structure of animal tissue on the one hand, and of medicinal substances on the other hand, it now seems to the observer that medical science at last has the opportunity and the facility to become an exact science.

"The Mass Spectrometer as an Industrial Tool" by A. O. Nier of the University of Minnesota. Ten years ago, the mass spectrometer was but a complex laboratory device understood and used by only a few specialists working in highly scientific fields. Today it is an important tool in any laboratory or plant where gas analyses are required or where vacuum problems are encountered. It has been used as a means of making continuous automatic gas analyses of the process gas used in the large diffusion plant for the separation of uranium 235. It has been used also as a tool for readily and rapidly locating vacuum leaks in plant equipment of many varieties. The mass spectrometer is a gas analyzer depending in operation upon electronic principles. The gas to be analyzed is ionized by electron impact, and the ions thus formed are separated by magnetic fields into beams according to the weights of the ions. From the relative intensities of the various ion currents, the composition of the gas may be deduced accurately.

"The Cathode-Ray Spectrograph" by R. Feldt and C. Berkley of the Dumont Laboratories. This instrument produces complete spectral distribution curves of the source under test, at the rate of 240 complete spectra per second. The spectrogram is presented visually on a cathode-ray tube. The instrument was developed initially to facilitate the study of microsecond current pulses on a cathode-ray tube fluorescent screen, but the color of any other varying light source may be studied similarly. Provision is made for the insertion of transmission samples, chemical samples, or filters for calibration. Continuous recording may be accomplished through the co-ordinated use of a camera attachment, and indications may be given in positions remote from the analyzer when desired. Possible applications include: instantaneous color matching, fluorescent lamp characteristics, spectrographic research, continuous process control, and transmission of spectral characteristics from remote or inaccessible locations.

RADIO AND RECORDING

"The Reduction of Background Noise in the Reproduction of Music From Records" by H. H. Scott (M '38) of the Technology Instrument Corporation. Conventional reproducing systems used with disk records or other noisy sources of audio-frequency signals have fixed transmission characteristics, regardless of the instantaneous requirements for reproduction of the applied signal with the optimum signal-to-noise ratio. By the use of "gate circuits" a system is achieved that is selective with respect to the type of applied signal, and which consequently tends to attenuate circuit types of signals while reproducing other types without substantial modification. Through proper design it is possible to obtain a system which will transmit with a high degree of fidelity those types of signals normally encountered in vocal and orchestral music, while at the same time discriminating against the usual types of background noises such as needle scratch and motor rumble. It also allows the reproduction of a greater frequency range than is common in conventional phonograph systems, thus reducing to a large degree the obvious difference between phonographic music and good frequency-modulation reception on a high grade radio phonograph combination. In its simplest form, the noise suppresser is reported as sufficiently inexpensive to be incorporated in a home-type phonograph, while the more elaborate version is suitable for use in broadcasting stations.

"Recent Developments in Magnetic Recording" by R. B. Vaile, Jr. (A '35) of the Armour Research Foundation. In the development of an ideal magnetic recorder some of the serious problems involved have to do with

1. The resolution of the recording head.
2. Distortion resulting from nonlinearity of the magnetization curve and other properties of the medium.
3. Constant speed drive.
4. Permanence of the record.
5. Transfer of the record from one part of the medium to an adjacent part.

The compromises which seem preferable in the present evaluation of solutions to these problems involve the use of

1. Longitudinal magnetization of the medium.
2. Speed between 0.5 and 5.0 feet per second.
3. Supersonic "bias."

In addition, extensive work has been done on the development of special materials, new test equipment, and special test methods.

MOBILE RADIO COMMUNICATION

This subject was discussed by a five-man panel, special attention being given to selective calling systems and to some of the specific items of equipment involved. "Signal Systems for Improving Railroad Safety" were discussed by K. W. Jarvis (M '34) consulting engineer of Winnetka, Ill.

NUCLEAR PHYSICS

"Some Fundamental Problems of Nuclear Power Plant Engineering" were considered by E. T. Neubauer of the Allis-Chalmers Manufacturing Company. As the temperature of an atomic reacting pile may range from a few degrees to millions of degrees, some of the important problems in the development of atomic power for commercial use pertain to the control of pile temperature and to the development of materials capable of withstanding ultrahigh temperatures. Another important problem is that of adequate shielding for the protection of personnel. On the basis of best present knowledge, regardless of the size of the pile, the total weight of shielding must be of the order of some 30 to 50 tons or more to suppress dangerous radioactive emanations. Incidental to any commercial development, therefore, it seems probable that most pile controls will be remote so as to keep operating personnel at safe distances. Further, indirect heat transfer methods must be devised to permit heat to be transferred from the pile to utilization or conversion equipment without a corresponding transfer of dangerous and undesirable radioactive emanations. The per-unit cost of the energy available today (about 1/10 of one per cent of the theoretically possible total) from fissionable material currently available is within reason and comparable with that of ordinary combustible fuels. The cost and complexity of conversion and utilization, however, impose other problems. It is to be expected that the use of atomic energy for power purposes probably will find its first application in naval vessels and then in commercial ocean-going vessels. Some observers consider that this is a possibility, if not a probability, within the next five years or so. According to the author "it may be possible that the further divisions of the nucleus will reveal that the building blocks of the atom are the electron and the positron, instead of the electron, positron, and neutron; if so, there is a small possibility that some day science will learn the art of directing the nuclear parts in some special electronic tube or conductor. It is worth a gamble, for if this can be discovered, the greatest possible use and the highest possible efficiency for nuclear energy will have been found."

"An Accelerator Column for Two to Six Million Volts" was presented by R. R. Machlett. The direct acceleration of electrons and positive ions to energies of several million volts, using the electrostatic constant-potential generator, has become an important physical technique with applications to nuclear research, deep X-ray therapy, and the X radiography of heavy metal sections. Multisection vacuum tubes in which charged particles are progressively accelerated and focused, have been developed for such application. These accelerator columns consist of a multitude of planary electrodes spaced three per inch and separated by glass insulation. For a five-megavolt tube the column is 110 inches long and contains 330 electrodes.

"The Betatron Accelerator Applied to Nuclear Physics" by E. E. Charlton (A '45) and G. C. Baldwin of General Electric Company described applications of another particle-accelerating device. Splitting or fission of the uranium atom by X rays takes place most effectively at energies of 20 million electron volts, but at higher energies processes other than simple fission occur, and their nature currently is being studied with the aid of the 100-million-volt betatron. The nature of these other processes of higher energies is inferred from studies with elements lighter than uranium. For example, a nitrogen atom disintegrated into hydrogen, helium, lithium, and several neutrons. Warning that further significant advances in the science of the atomic nucleus will depend upon discoveries in fields not yet explored, Doctor Baldwin stated that "physicists today are hardly nearer an understanding of the atomic nucleus and of the nature of the particles which compose it than they were several years ago, even though the successful development of the atomic bomb may lead one to think that the fundamental questions of nuclear science all are answered. On the one hand, the radiation or absorption of energy by the electrons in a radio antenna, the flow of electrons in a vacuum tube, and the motion of electrons in an atom all can be described with high precision by well-established physical theories based upon simple laws of the electromagnetic fields of elementary charges of electricity. On the other hand, many nuclear phenomena are an almost unsolved riddle. The neutrons and protons which compose a nucleus are held together by extremely strong forces of very short range. We do not know the laws by which these forces act. We do not know what relationships exist, if any, between the nuclear and the electromagnetic fields. We know very little about the elementary particles of the nature of themselves—why there are so many kinds of them, how they interact with each other, whether they have internal structure, why some of them are radioactive, and so forth."

This 5-kw magnetron operates at 1,050 megacycles and is a General Electric Company product

INSTITUTE ACTIVITIES

Winter Meeting Papers Presented in Four Main Groupings

An all-inclusive professional program and diversified social activities are promised for the AIEE winter meeting, January 27–31, 1947, in New York, N. Y. On the social side, as usual, a smoker and dinner dance will be held. Theater tickets will be available for out-of-town members, and special entertainment for women will be provided. Inspection trips will be arranged to places of interest and near-by industries. On the professional side technical sessions and conferences, segregated in four broad groups—industry, power, communication and science, and general applications—will cover practically the entire field of electrical engineering. The general session will be the common meeting ground of all four groups. During the meeting, presentation of the Edison, John Fritz, and Hoover Medals will be made. Details of the awards will be announced later. Meeting headquarters will be in the Engineering Societies' Building.

INDUSTRY GROUP

In the industrial field conferences will be held to highlight problems in the textile industry and to discuss the optimum industrial voltages for distribution systems. A session on industrial instrumentation is in prospect, and still another session will present a variety of industrial power applications.

For the electric welding people a conference will be arranged to discuss the application of resistance welders and the problems of power supply. A session on resistance welding will deal with battery welding, the secondary impedance of resistance welders, and calculations on the effect of introducing ferromagnetic welding materials into the throat of resistance welders, all of which are factors affecting the design of these machines.

Metallic rectifiers, which have come into extended use in many industries since the war, will be the subject of a conference. The proposed standards and the proposed test code for metallic rectifiers will be presented, together with other papers on this subject, and interested groups in the Army, Navy, and Air Corps will be invited to attend.

Industrial control devices will be the subject of another session, and papers will be presented on the design details of components, as well as on applications circuits of complete controllers. Also of interest to industrial engineers are a session in the communication and science group, dealing with the theoretical aspects of servomechanisms and a conference on electronic instruments.

POWER GROUP

A timely session is being held to revive interest in hydroelectric systems. A conference on interconnected systems is being arranged, as new problems have arisen in the five-year interlude since the last session on this subject. Considerable progress has been made in the solution of these problems, which makes the subject pertinent at this time.

In the field of power transmission and distribution, sessions are in prospect to present the latest information on lightning surges, grounding, and protective apparatus. In a session on electric machinery consideration will be given to the design of capacitor motors, the new NEMA standards for fractional horsepower motors, as well as the effect of steep front waves and chopped waves on transformers, and short-circuit standards for three-winding transformers.

COMMUNICATION AND SCIENCE

In the field of communication three sessions and a technical conference will be arranged. One of the sessions will deal with rural telephone facilities. The problem of providing telephone service to more people living in rural areas is now being attacked vigorously from many angles. Papers will be presented on each of the many approaches being used. Included will be papers on the design and application of carrier telephone systems for operation over rural power lines, the joint use of pole lines for power and telephone service, new methods of telephone wire construction for rural areas, and the use of radiotelephone to provide rural service.

Another session will be devoted to television. With more television stations going on the air and television receivers reaching the market in ever greater numbers, television is in the forefront of engineering advances and many aspects of its progress will be presented at this session. Included will be papers on transmitter and camera equipment, television receiver design, color television, and networks for television, both coaxial cable and microwave radio.

The session on communications components and techniques has scheduled papers on a variety of interesting subjects, for example, papers on recent improvements in telephone transmission systems, modern ocean cable technique, new developments in telephone switching systems, and subjects related to magnetic recording.

The technical conference in this group will be on testing telephone receivers and transmitters. Brief talks will be given by representatives of many concerns and institutions interested in this matter, and provision will be made for discussions and questions from the floor.

Applications in several fields of engineering will be emphasized at the session on instruments and measurements. Papers on new moving magnetic instrument design and magnetometer design for measurement of magnetic fields will stress wartime techniques as applied to peacetime uses. Other papers describe compensation for transformer loss in metering, measurement of crest voltages in ignition systems, and spectrographic measurements of light in rapidly changing spectra by use of a cathode-ray oscillograph for control or measurement purposes.

The conference on electronic instruments will consider a variety of applications, such as an electronic null detector for use with impedance bridges, pulse echo measurements on telephone and television facilities, defect location of materials by supersonic waves, and a general review of the electronics measuring field.

In the field of servomechanisms another session will attempt to answer some of the technical questions raised during the summer meeting in Detroit last June and to present more of the recent advances. A conference on servomechanisms also is proposed for the purpose of bringing together representatives of the technical personnel working in this field and representatives of the industrial people. An effort will be made to bridge the gap between the point of view of the technical man and the point of view of the nontechnical user of servomechanisms.

In the field of basic sciences a conference on applied mathematics has for its purpose the bringing together of mathematicians and engineers to discuss different engineering problems which have arisen during the past few years concurrently with new developments in the engineering and scientific fields. A session on topics of wide appeal in the basic science field will be held, and a conference is planned on the subject of the use of large-scale computing devices.

Many papers will be presented in various sessions on applications of electronics. In addition, a session on electronics will present a group of papers on rectifiers.

GENERAL APPLICATIONS

In the field of land transportation three sessions will be devoted to broad aspects of the subject. More specifically a session on light traction will deal mainly with the rapid transit system and subway equipment for New York, N. Y. Major General Charles P. Gross, chairman of the Board of Transportation, will give a special address at this session. Another session on electric railways will consider railroad electrification, power conversion and transmission costs, current collectors for high-speed service, and the modern Presidents' Conference Committee street car. Recent developments in various phases of Diesel-electric locomotives also will occupy a session.

Aircraft safety will be the subject of an all-day conference to discuss the factors affecting the safe operation of electric apparatus. Discussions will be prepared on assigned topics divided among the air line operators, the air frame manufacturers, and the accessory manufacturers. The program will be divided in six panels, as follows: safety, history, and requirements; types of electrical failures; protection in the electric system; application considerations; the safety program; and general discussion. The entire program will be conducted from a noncritical point of view with an attempt to bring out only constructive contributions.

The committee on safety will sponsor a symposium on grounding of portable equipment prepared by the committee on article 250, grounding, of the electrical committee of the National Fire Protection Association, for consideration in the development of the National Electrical Code; and discussions of the report will be presented by producers and users of portable electric equipment and by producers, installers, and users of the connecting cord and plug equipment involved in such grounding.

The enormous increase in the use of portable electric tools and other equipment in manufacturing establishments makes it highly important that engineers concerned with the development, use, and maintenance of such equipment be well informed as to the need for safeguarding and methods of accomplishing it.

Quality control will be discussed in a conference jointly sponsored by the American Society for Quality Control and the Institute's subcommittee on applications of statistical methods. The general theme of the conference will be applications to consumer goods. G. D. Edwards (M'26) president of the American Society for Quality Control, will speak on the purposes and progress of the society. This will be followed by two papers dealing with special applications of quality control to consumer goods. The conference will close with a general discussion, including all aspects, with particular reference to certification of consumer goods.

Further details of the entire program will be announced in the January issue of *Electrical Engineering*.

New AIEE Publication Policy Becomes Effective in January 1947

By Charles F. Wagner, Chairman AIEE Publication Committee

One of the fundamental purposes of the Institute is the dissemination to its members of information concerning new theories, new apparatus, and new developments throughout the electrical field. Incidental to its responsibility under this objective, and pursuant to action of the board of directors (*EE, June 1946, p 270*) the AIEE publication committee has been developing ways and means of improving the Institute's publications policy and procedure with the objectives of better serving the greatly enlarged and rapidly growing membership and of covering adequately the active and ever-growing electrical field.

The purpose of this report is to describe and explain the modifications in present publication policy and procedure constituting the new AIEE publication policy, which is to become effective immediately with 1947 publications. Before discussing the new policy, it is desirable to review the present policy which now falls somewhat short of membership needs, and to define the various kinds of publication materials that come to the Institute for consideration.

PRESENT PUBLICATION PLAN

The present AIEE publication policy was established in 1940. Most of the material published by the Institute is submitted as technical papers. These papers are referred first to the technical program committee for review and approval; if approved, they then are scheduled for presentation at a national or District meeting, where they are open for discussion.

Under the present publication policy, all technical papers so approved and recommended by the technical program committee are published in the AIEE *Transactions*, which is issued annually covering a calendar year and which serves as the permanent repository of material covering the progress and development of the electrical art. A *Transactions* section of the monthly magazine *Electrical Engineering* is set aside and allocated for publication of approved technical papers, this being in actuality a monthly preprint of identical pages which will constitute the current annual *Transactions* volume. Because of cost limitations, space has not been available in the monthly magazine for all approved papers. The overflow papers are printed twice yearly, in June and in December, in a special publication known as the "Supplement to *Electrical Engineering—Transactions* section." These supplements are available to Institute members at 50 cents each or $1 per year. Discussions of technical papers do not appear in *Electrical Engineering*, but first become available to the membership in the supplements, after a lapse of several months; for example, winter meeting discussions in the June supplement. Thus a member automatically receives most of the technical papers in the *Transactions* sections of *Electrical Engineering*, and by purchasing the supplements receives all of the remaining approved technical papers and all approved pertinent discussion. All this material appears in the annual AIEE *Transactions*. Discussions, however, do not appear on *Transactions* pages immediately adjoining the paper to which they apply, but appear elsewhere in the volume.

Certain of the papers submitted by contributors to the technical program committee and the technical committees for review and consideration are adjudged to be appropriate and timely for presentation and discussion at a national or District meeting, although they are deemed to be inappropriate for permanent record. Although these papers are presented at meetings, they are not published in the *Transactions* sections of *Electrical Engineering*, the supplements, or the *Transactions*. In current parlance they are known as "ACO" (advance copies only) papers; formally, this series of papers is designated as "AIEE miscellaneous papers" to distinguish them from the regular "AIEE technical papers" series.

To facilitate and to promote discussion of papers, individual pamphlet copies of each technical paper (regular or ACO) are issued in advance of the meetings at which the papers are scheduled to be presented and discussed. These pamphlets are known as "advance copies" or "preprints" of the technical papers. As the time element is of essential importance in relation to the meetings programs and speed is required in the physical reproduction of these papers to make them available to interested members as far as possible in advance of their presentation at meetings, a quick method of reproduction is resorted to that is known as the photolithographic or "photo-offset" process. In this process of reproduction, the author's original manuscript copy and illustrations are reproduced photographically approximately in facsimile, and without any of the intervening steps necessary for regular printing. The photolithographic process is a more expensive reproduction process than printing where substantial quantities are involved, but does offer the advantage of speed in reproduction. These "advance copies" or "preprints" therefore are produced only in limited quantities, and are made available to members approximately at cost.

To facilitate the efforts of District meeting committees in preparing programs best adapted to the timely and particular interests of the region to be served by District meeting, another class of paper has been given a recognized status by the Institute. This class of paper, known officially as "District papers" (DP) is authorized for acceptance for presentation and discussion at a District meeting solely at the discretion of the District meeting committee and without the requirement of advance review by the technical program committee or the technical committees. Correspondingly, such papers have status only as parts of the local District meetings programs, where they may be discussed along with other papers, but carry with them no commitment or obligation for publication of either paper or discussion. In some instances, of course, the subject matter of District papers may be developed further and processed through the regular AIEE channels previously discussed and ultimately become eligible for inclusion in *Transactions.* These District papers normally are not prepared in pamphlet form by the Institute, nor are they generally available even at the meetings where they are scheduled for presentation, except in instances where authors may provide mimeographed or other copies.

Still another class of paper is involved currently in national and District meetings programs, known as "conference papers" (CP). These conference papers in a great many instances are not papers at all, but informal oral presentations made at conference sessions or round-table discussions which are organized from time to time on subjects of current interest under the encouragement of the technical program committee and the direct sponsorship of one or more of the technical committees. Like the previously mentioned District papers, conference papers are wholly informal presentations, not reproduced or distributed by the Institute, and carry no commitment for publication of paper or discussion. Such papers may be processed subsequently through regular channels and ultimately become *Transactions* material. Authors of conference papers sometimes make a limited number of mimeographed or other copies available at their own expense for use at the meetings. Otherwise, to preserve the complete informality of the conference session, there is no record of what transpires at such meetings or conferences, except as it may be covered in news reports in *Electrical Engineering.*

Electrical Engineering, under the present publication policy, contains many engineering and other articles, news items, personals, and other material of current general interest to AIEE members in addition to the *Transactions* section previously mentioned. This material is drawn from AIEE and other sources. The choice of material, as well as the general management and publication of the magazine, is the responsibility of the editor.

CURRENT PROBLEM

The electrical art has become so broad in scope and so specialized in nature that any particular technical program paper now is of interest to only a relatively small portion of the entire membership. Thus, under the present publication plan the average member is receiving many specialized papers in fields of relatively little interest to him. This repeatedly has been brought to the attention of the publication committee with the corollary suggestion that in most such instances an abstract or digest presenting on an informational basis the essence of such specialized papers would better satisfy the average member's requirement, leaving the full technical paper for the specialist. Under the present publication policy, the space required to print these technical papers in full in *Electrical Engineering* crowds out material of broader general interest, a situation that is growing more acute as the national and District meeting activities grow and the number of technical papers increases. The policy requiring publication of these papers in full in *Electrical Engineering* places undesirable restrictions and limitations upon the editor in his efforts, under his assigned responsibility, to produce in *Electrical Engineering* a magazine of greater service and more general appeal to the membership at large.

Discussions and rebuttals of technical papers, under the present publication plan, appear as already noted only in the *Transactions* and in the "Supplement to *Electrical Engineering—Transactions* section," and in any event only after the lapse of so long a time that much of the interest in the discussion is lost. Further, such discussions do not appear on pages immediately adjoining the papers to which they apply, thus necessitating reference to two different locations to obtain the complete story concerning any particular paper. The publication committee has received continual criticism of this practice.

NEW PUBLICATION PLAN

In the new publication plan, the *Transactions* section will be eliminated from *Electrical Engineering* and in its stead will be established a new series of publications to be known as "AIEE Proceedings" which will consist principally of all technical papers with their discussions, and which also may contain other appropriate material from national and District meetings. Individual members will be given the opportunity to select designated groups of these technical papers to be received free of charge, and other papers to be received as desired at a nominal charge to cover cost. More specifically the publications and policy of the Institute may be described as follows:

1. *Electrical Engineering.* Coincidental with the elimination of the *Transactions* section, the general-interest content of *Electrical Engineering* will be expanded, enhanced, and given greater flexibility to satisfy more fully the current and future requirements of the membership. It will be characterized especially by the following attributes.

(*a*). Subject only to general policy, the management and publication of *Electrical Engineering* will be entirely the responsibility of the editor.

(*b*). The general-interest content will be approximately doubled, contemplated to be at least 100 pages per month.

(*c*). The contents will be divided broadly into two general categories: (1) general articles, to be drawn from all currently available sources including material submitted for technical program papers after, but regardless of, action by the technical program committee, selected on the basis of timeliness and general-interest value to the membership at large and treated accordingly; (2) news material, comprehensively and appropriately reflecting all AIEE and related activities, also appropriately reflecting other current activities of interest and significance to the membership. For example, with reference to AIEE programs, whether technical sessions or conferences, it is contemplated that all technical program papers and all other appropriate material will be reflected currently in *Electrical Engineering* for the purpose of keeping the membership fully and currently informed.

(*d*). Summary abstracts of technical papers will continue to be included as at present and will be published as far as possible in advance of the meeting at which the papers are scheduled for presentation and discussion.

(*e*). The weight and quality of both inside and cover paper stock will be improved. This will recover the losses sustained incidental to United States Government wartime regulations and policies, and should obviate the justifiable criticisms that have been received from readers and advertisers concerning the unsatisfactory quality of reproduction of illustrations and other matter.

(*f*). A more aggressive advertising policy will be inaugurated to capitalize more fully upon the established prestige and substantial standing of *Electrical Engineering.*

(*g*). Distribution will continue to be made automatically to the entire AIEE membership; to be available also to nonmembers at an appropriate subscription price in accordance with postal regulations and as provided in AIEE bylaw 103 and 104.

2. *Preprints.* To encourage and facilitate discussion of technical papers at national and District meetings, the present practice of issuing advance pamphlet copies or "preprints," of technical program papers by the photolithographic process or an appropriate equivalent will be continued. Distribution of this material is to be made as follows.

(*a*). The minimum necessary number of copies will be available for distribution without charge by or at the specific and detailed request of the committee responsible for the development and approval of the manuscript in question. Such distribution is intended to be made prior to the meeting at which the paper is scheduled for presentation and discussion, and primarily for the purpose of stimulating qualified discussion.

(*b*). Copies will be available upon request to the entire AIEE membership at cost, such cost to include all production, overhead, and distribution costs; available to others at 100 per cent above members price. Following is the schedule of members prices:

Size of Pamphlet (Pages)	At Registration Desk	By Mail
12 or less	$0.10	$0.15
16	.15	.20
20	.20	.25
24 or more	.25	.30

3. *Proceedings.* For the purpose of better serving those among the membership who are especially interested in the full and

complete technical and mathematical details of subject matter within the field of their particular professional interest, a new series of publications is to be established, to be known as "AIEE *Proceedings*." The *Proceedings* will contain the full text of all papers approved by the technical program committee for publication in *Transactions* and for presentation at national or District meetings together with all discussions of those papers as approved by the technical program committee; also, may contain "miscellaneous (ACO) papers" approved by the technical program committee for presentation at national or District meetings; and other appropriate material. The *Proceedings* thus is planned to contain all material now published in the *Transactions* sections of the 12 monthly issues of *Electrical Engineering* and the two semiannual supplements, and in addition a very considerable amount of meeting material heretofore not published.

To provide maximum flexibility in production and distribution and thereby expedite materially the distribution of such material, and in deference to the long expressed wishes of the technical specialists among the Institute's membership who are most interested in the full and complete details of technical papers and their related discussions, the *Proceedings* will be issued in the form of individual printed pamphlets. Each pamphlet will contain the full text of an approved technical paper, and all approved pertinent discussion. The factors which will control the date of completion and initial distribution of this material will be (*a*) the determination by the technical program committee that no discussion is available or (*b*) the release by the technical program committee of discussion which it has approved for publication. On this basis, this material will be available to those who especially want it several months earlier than is possible under the present publication plan.

Each Fellow, Member, and Associate of the AIEE will be given an opportunity to select for receipt free of charge a limited quantity of the total number of *Proceedings* sections (pamphlets), the quantity designed to be sufficiently generous to enable the average member to obtain those papers in which he is most intimately concerned. Others will be available at cost. *Proceedings* sections also will be available at cost to Student Members and *Electrical Engineering* subscribers.

4. *Annual Transactions.* Current policies and procedures with reference to AIEE *Transactions* are to be continued without change, except that the flexibility introduced by the previously discussed changes in publication procedure will enable discussions to be assembled with the papers to which they apply. Bound volumes will continue to be issued on the basis of material presented during a calendar year, in single or multiple volumes, depending upon the amount of material and other considerations. *Transactions* will contain all formal AIEE technical papers and related discussion as approved by the AIEE technical program committee, and other record material appropriately approved. Distribution will continue on the basis of advance order; to the AIEE membership at cost, to others at appropriately higher prices.

5. *Special Publications.* To provide a facility in a field of publication services heretofore not satisfied adequately by the Institute, a series of "AIEE Special Publications" will be established and its use encouraged. It is contemplated that such publications will be issued from time to time and in appropriate relation to other AIEE publications to accommodate material of especial value and importance in its field but of limited general interest or applicability, or material peculiarly convenient and appropriate in the form of separate and substantially bound books or booklets. It is contemplated that these special publications may accommodate material highly technical or mathematical in content, or specialized in subject matter, such as the "AIEE Lightning Reference Book," the AIEE Joint Subcommittee report "Telemetering, Supervisory Control, and Associated Circuits," and the "Bibliography on Automatic Stations, 1930–1941," and other bibliographies, consolidated reprints, special reports, surveys, and similar material.

The format, typographical style, paper, and printing will be of a quality appropriate to an AIEE publication and commensurate with their positions in the AIEE family of publications. Special publications are to be issued on an entirely self-supporting basis, the proceeds from sales in effect constituting a revolving fund for continuation and stimulation of the project.

Distribution of special publications is to be on the basis of individual order, at cost to the AIEE membership; to others at 100 per cent above the prices which will be set for members.

6. *Reprint Service.* As a special service which has proved to meet many membership needs, the present practice of publishing quantities of reprints or preprints of AIEE published material will be continued. Such reprints usually will be produced on the basis of specific orders, but occasionally upon the basis of anticipated need. Such reprint services are to be performed on an entirely self-supporting basis.

7. *Annual Index.* To provide a consolidated single authoritative reference guide to all AIEE material published during any calendar year, a complete composite AIEE annual index is planned, to cover appropriately the content of *Electrical Engineering*, *Proceedings*, *Transactions*, and special publications. This index will continue the basic alphabetical multientry subject listing system originally developed by *Electrical Engineering* and used with eminent success over a period of many years. Distribution will be automatic and without charge to the AIEE membership and to regular subscribers to AIEE publications; upon request to others.

8. *Yearbook.* Publication of the Yearbook is to be continued as at present. Distribution is on the basis of individual requests.

9. *Conference Papers.* The informal AIEE conference papers will be treated very much as in the past. As such, they do not have publication status, although the editor of *Electrical Engineering* specifically is authorized to make appropriate use of conference material subject to the approval of the authors involved.

Great Lakes District Meets in Indianapolis

Sponsored jointly by the Fort Wayne and the Central Indiana Sections, the seventh Great Lakes District meeting was held in Indianapolis, Ind., October 9–11, 1946, with headquarters at the Claypool Hotel. Most of the sessions were held at the hotel the others were held in the auditorium of the Indiana War Memorial. The verified attendance was 298; other attendance details and comparisons with previous meetings of the Great Lakes District are given in accompanying tabulations.

The history of this meeting reveals quite a checkered career for a District meeting. Originally planned and scheduled to be held in Fort Wayne in April 1941, the meeting was indefinitely postponed in the war aftermath of Pearl Harbor. Picking up postwar where it had left off in 1941, the Fort Wayne Section had completed all plans and preliminary announcements to hold the meeting in Fort Wayne this fall, only to be overtaken by the hotel situation wherein Fort Wayne hotels withdrew the commitments of room and meeting space necessary. It was into this breach that the Central Indiana Section stepped, to provide the necessary hotel and other physical arrangements for the meeting in Indianapolis, and to join wholeheartedly with the Fort Wayne Section in carrying the meeting out to a very effective conclusion. Even the heavens collaborated, staging a once-in-a-century display shower of meteors on the opening night of the meeting.

CONFERENCE TYPE OF MEETING

A noteworthy feature was the fact that only four regular technical program papers were presented, the other 26 being informal presentations classed as "District papers" or "conference papers." This type of program proved to be both popular and effective in presenting for informal exchange and discussion a wide variety of up-to-the-minute topics which catered effectively to a very wide range of member interest. In fact, it drew a registered attendance of more than 80 nonmembers and many additional who did not register.

The 30 subjects were divided among six conference sessions covering electric machinery, electronics, power transmission and distribution, industrial power applications, communications, industrial control, and basic sciences. In general, the sessions ran two in parallel without too much conflict in member interest, the loss from conflict and interest being more than offset in the opinion of most observers by the additional scope of subject matter that could be accommodated by parallel sessions.

A more detailed summary report is given in another item elsewhere in these pages, covering the various technical topics discussed.

PROFESSIONAL ACTIVITIES DISCUSSED

One entire afternoon was devoted to a general conference on "Institute Activities and Study of the Organization of the Engineering Profession," under the chairmanship of AIEE Vice-President T. G. LeClair (F '40) of the Great Lakes District who also is chairman of the professional activities subcommittee of the AIEE committee on planning and co-ordination.

President J. Elmer Housley (F '43) discussed "How the Engineer Should Look at Tomorrow," stressing the need for engineers to give attention to problems of public and professional concern as well as to technical topics. He stressed the need for the AIEE to give steadily broader and better service to its membership, and briefly mentioned the efforts currently being made to accomplish this, and to learn more accurately what the membership wants by conducting discussions with representative groups throughout the country. He urged particularly much more active attention to industrial applications and problems, special efforts to develop the interest and greater participation of the executive and administrative group of electrical engineers, who deal primarily with management and labor relations problems rather than the more technical subjects; he urged especial effort to develop a better degree of co-ordination between the executive group and the younger engineers.

Vice-President LeClair presented a summary report covering the various projects that have been and are undergoing study by the committee on planning and co-ordination. As a preliminary to the specific discussion for which the session was planned originally, he also presented a comprehensive review of the study currently being made by his subcommittee on professional activities covering organization of the engineering profession, reflecting the high lights of the full report originally published in the April 1946 issue of *Electrical Engineering*, pages 169–73. Most of the discussions presented were from representatives of the various sections in the District, in response to previous solicitation. A few spontaneous comments were forthcoming from among the 12 persons in attendance including one from a visiting mechanical engineer who enthusiastically urged that "The engineers must organize as a profession or be organized by the government as a trade." In urging a single society for all engineers, one commentator made the broad assertion that "All the national societies are dominated by employers and hence are opposed to the over-all organization of individual engineers," but did not elaborate upon or substantiate this assertion. In general, the various discussers paid some homage to the theoretical idealism of the single-society idea, but as a matter of fact a substantial majority of all the commentators recommended "Plan *C*, Federation or Council of Existing Engineering Societies" and urged some early action in that direction.

Comparison of Great Lakes District Meeting Attendance

1946	October 9–11	Indianapolis, Ind.	298
1939	September 27–29	Minneapolis, Minn.	230
1935	October 24–25	West Lafayette, Ind.	481
1932	March 14–16	Milwaukee, Wis.	552
1929	December 2–4	Chicago, Ill.	600
1927	November 28–30	Chicago, Ill.	900
1926	May 6–7	Madison, Wis.	180

Analysis of Registration at Great Lakes District Meeting

Classification	Central Indiana Section	Fort Wayne Section	District 5*	Other Districts	Total
Members	68	19	74	31	192
Student Members	10				10
Men guests	50	13	15	5	83
Women guests	4	5	2	2	13
Totals	132	37	91	38	298

*Outside Central Indiana and Fort Wayne Sections.

ENTERTAINMENT AND TRIPS

The principal feature of the District meeting entertainment activities was the general banquet held Thursday evening, October 10, in the James Witcomb Riley room of the Claypool Hotel, with Vice-President T. C. LeClair presiding. Guests were treated to an excellent and lucid lecture on the "Nature and Implications of Atomic Energy," by Doctor Frank H. Spedding of Iowa State College, Ames. Doctor Spedding was in the midst of wartime atomic research from its earliest beginnings, and currently is a member of the Manhattan District Declassification Committee (MDDC), the group upon whose shoulders rests the responsibility of recommending those portions of the scientific information developed in wartime atomic research that can be released at this time to aid in scientific, technical, and medical research without impairing national security. While emphasizing the enormous potentialities of nuclear and atomic energy, for constructive use or for destruction, Doctor Spedding thoroughly debunked the sensational newspaper, magazine, and book accounts of quasi scientists and others which have caused many persons to fear that the earth itself was in dire danger of physical destruction. Doctor Spedding expressed the opinion that atomic power plants probably would become a reality within something like five years or so, especially for large naval and merchant ships, where the space required for fuel and power plants are of especially critical importance and economic significance.

To avoid interference with conference sessions, and at the same time to facilitate transportation and other arrangements, all technical inspection trips were scheduled for Friday afternoon after the last technical conference session. These trips provided for visits to the Marmon Herrington Company, manufacturers of mobile transporation equipment including all-electric trolley coaches; to the Esterline-Angus Company, Inc., makers of recording instruments; to the Eli Lilly and Company plant and laboratory, devoted to medicinal products; to various local power facilities of the Indianapolis Power and Light Company and of the Public Service Company of Indiana; to several of the plants of P. R. Mallory and Company, manufacturers of various radio controls and equipment; to the Chinook Mine of the Ayrshire Collieries Corporation at Staunton, Ind., and to the Maumee Collieries Corporation mine at Jasonville, Ind., two large strip-mining operations utilizing large electric shovel equipments; and to Purdue University, West Lafayette, Ind.

Activities especially aranged for women guests included a trip to the James Witcomb Riley Hospital for children, one of the outstanding hospitals for the medical care of children; a trip to the John Herron Art Institute; a tea and fashion show at the L. S. Ayres and Company, department store; the District meeting banquet.

Equipment exhibits originally had been planned in connection with the meeting, but did not materialize because of industrial production conditions.

COMMITTEES

The joint sponsorship of the Great Lakes District meeting in Indianapolis by the Fort Wayne Section and the Central Indiana Section is reflected in the committee organization. In general, Fort Wayne carried the responsibility for the development of the technical program and the Central Indiana Section carried the responsibility for arranging and managing the meeting activities. The personnel responsible for the success of the meeting include the following

General Cochairmen

F. H. Fleischer
S. C. Leibing

Meetings and Papers Committee

C. M. Summers, *chairman*
M. A. Baker
F. O. Nottingham
M. L. Schmidt
W. K. Self
E. S. Sullivan

Inspection Trips

C. E. Chatfield, *chairman*
H. A. Berkheiser
L. E. Briscoe
W. A. Gentry
C. E. Parks
C. R. Swenson
John Webb

Hotels, Registration, and Entertainment

S. C. Leibing, *chairman*
L. T. Clipson
E. G. Downie
R. B. Hull
G. W. Kintner
C. E. Parks
J. R. Pies
H. F. Roempke
John Sears
Fred Stout
M. F. Schonefeld
I. W. Strong
C. R. Swenson
R. O. Whitesell

Finance Committee
E. A. Linke, *chairman* W. H. Bollinger

Publicity Committee
J. L. Wright, *cochairman* S. A. Zimmerman, *cochairman*

Student Activities Committee
S. W. Winje *chairman*

Assistance in the actual handling of registration and related details was provided by the Indianapolis Convention and Visitors Bureau.

Contents Announced for December 1946 Supplement

Technical papers presented before AIEE technical meetings during 1946 and not published in the monthly *Transactions* section of *Electrical Engineering* or in the June supplement will appear in the December 1946 "Supplement to Electrical Engineering—Transactions Section." The December supplement will contain 30 technical papers, discussions of those papers, and discussions of the papers already published in the July–December monthly sections. Issuance of the December supplement will complete publication of the papers and discussions presented before the 1946 North Eastern District meeting in Buffalo, N. Y., April 24–25; the 1946 summer meeting, Detroit, Mich., June 24–28; the 1946 Pacific Coast meeting, Seattle, Wash., August 26–30; and the 1946 Great Lakes District meeting, Indianapolis, Ind., October 9–11.

In order that distribution of copies of the supplement can be made in February 1947, those desiring to receive copies should enter advance orders promptly.

Papers appearing in the December 1946 supplement, abstracts of which have been published in *Electrical Engineering* in advance of the meetings are:

Air Transportation

46-129—Electronic Control of Frequency Converter Sets for Testing Aircraft Models; *G. W. Heumann (A '44), W. F. Strong (A '41).* Abstracted in the June 1946 issue, page 278.

Basic Sciences

46-119—Formulas for Calculating the Inductance of Channels Located Back to Back; *W. G. Schwantz (Student), T. J. Higgins (M '46).* Abstracted in the June 1946 issue, page 279.

46-120—Negative Resistance Effects in Saturable Reactor Circuits; *J. M. Manley, E. Peterson (M '26).* Abstracted in the June 1946 issue, page 279.

46-195—The Use of Business Machines in Determining the Distribution of Load and Reactive Components on Power Line Networks; *P. D. Jennings, G. E. Quinan (F '18).* Abstracted in the August–September 1946 issue, page 411.

Communication

46-87—Radiotelemetering for Testing Aircraft in Flight; *C. L. Frederick (M '46).* Abstracted in the April 1946 issue, page 174.

46-183—Application of Compandors to Telephone Circuits; *C. W. Carter, Jr. (M '39), A. C. Dickieson (M '41), D. Mitchell.* Abstracted in the August–September 1946 issue, page 411.

Domestic and Commercial Applications

46-184—Specific Engineering Problems in Rural Electrification and Electroagriculture; *M. M. Samuels (F '24).* Abstracted in the August–September 1946 issue, page 412.

Future AIEE Meetings

Winter Meeting
New York, N. Y., January 27–31, 1947

North Eastern District Meeting
Worcester, Mass., April 23–25, 1947

Summer Meeting
Montreal, Quebec, Canada, June 9–13, 1947

Middle Eastern District Meeting
Dayton, Ohio, September 23–25, 1947

Midwest Meeting
Chicago, Ill., November 3–7, 1947

Electric Machinery

46-116—Electromechanical Transient Performance of Induction Motors; *C. N. Weygandt (A '37), S. Charp (A '42).* Abstracted in the June 1946 issue, page 280.

46-136—Current and Torque of D-C Machines on Short Circuit; *T. M. Linville (M '34).* Abstracted in the June 1946 issue, page 280.

46-164—Oscillations of a High-Voltage Secondary Winding; *A. Boyajian (F '25).* Abstracted in the June 1946 issue, page 280.

46165—Temperature Limits for Short-Time Overloads for Oil-Insulated Neutral Grounding Reactors and Transformers—II; *V. M. Montsinger (F '29), J. E. Clem (F '38).* Abstracted in the June 1946 issue, Page 281.

46-167—Rotating Electric Machine Time Constants at Low Speeds; *C. Concordia (M '37).* Abstracted in the June 1946 issue, page 281.

46-189—Dielectric Strength of Station and Line Insulation to Switching Surges; *P. L. Bellaschi (F '40), L. B. Rademacher (A '45).* Abstracted in the August–September issue, page 412.

46-202—Asynchronous and Single-Phase Operation of Synchronous Machines; *A. W. Rankin (M '45).* Abstracted in the October 1946 issue, page 472.

Induction and Dielectric Heating

46-124—Design of Induction-Heating Coils for Cylindrical Magnetic Loads; *J. T. Vaughan (A '44), J. W. Williamson (A '43).* Abstracted in the June 1946 issue, page 282.

Industrial Power Applications

46-194—Self-Excited Electromagnetic Drive for a Resonant Fatigue Machine; *A. R. Wilson (M '35).* Abstracted in the August–September 1946 issue, page 412.

Instruments and Measurements

46-133—Functional Analysis of Measurements; *I. F. Kinnard (F '43).* Abstracted in the June 1946 issue, page 283.

46-143—High-Vacuum Leak Testing With the Mass Spectrometer; *W. G. Worcester (A '42), E. G. Doughty.* Abstracted in the June 1946 issue, page 283.

46-181—An Oscillograph for Recording Transient Recovery Voltages; *W. G. Hoover (A '34).* Abstracted in the August–September 1946 issue, page 412.

Land Transportation

46-186—Comparisons of Railway Motive Power for Operations in the Pacific Northwest; *T. M. C. Martin (A '45).* Abstracted in the August–September 1946 issue, page 412.

Power Generation

46-108—Tennessee Valley Authority Hydroelectric Stations—Electrical and Mechanical Design; *R. A. Hopkins (F '43), H. J. Peterson.* Abstracted in the June 1946 issue, page 283.

46-144—Improvements in Performance of Hydroelectric Generating Units on the 2,000,000-Horsepower Saguenay System; *F. L. Lawton (M '36).* Abstracted in the June 1946 issue, page 284.

46-152—Modern Excitation Systems for Large Synchronous Machines; *J. B. McClure (A '29), S. I. Whittlesey (A '45), M. E. Hartman.* Abstracted in the June 1946 issue, page 284.

46-154—The Development of Modern Excitation Systems for Synchronous Condensers and Generators; *F. M. Porter (M '45), J. H. Kinghorn (M '46).* Abstracted in the June 1946 issue, page 284.

46-182—Long-Distance Power Transmission as Influenced by Excitation Systems; *C. Concordia (M '37), S. B. Crary (F '45), F. J. Maginniss (A '43).* Abstracted in the August–September 1946 issue, page 413.

46-117—The Electrical Performance of Ceramic Dielectrics at Elevated Temperatures; *H. A. Frey (M '43), J. A. Jesatko (A '41).* Abstracted in the June 1946 issue, page 284.

Power Transmission and Distribution

46-196—Characteristics of a 400-Mile 230-Kv Series-Capacitor-Compensated Transmission System; *B. V. Hoard (M '36).* Abstracted in the August–September issue, page 413.

Protective Devices

46-170—Performance Criteria for Current Limiting Power Fuses—I; *E. W. Boehne (F '43), C. L. Schuck (A '30).* Abstracted in the June 1946 issue, page 285.

46-171—Performance Criteria for Current Limiting Power Fuses—II; *E. W. Boehne (F '43), C. L. Schuck (A '30).* Abstracted in the June 1946 issue, page 286.

46-173—Simplicity in Protective Relaying; *L. F. Hunt (F '38).* Abstracted in the June 1946 issue, page 286.

AIEE Survey on Vacuum Tubes Progresses

The survey of instrument manufacturers, in the interest of providing more reliable vacuum tubes for use in electronic instruments, is well under way, W. R. Clark (M '44) chairman of the joint subcommittee on electronic instruments, has announced.

Some of the items covered in the questionnaire, which has been mailed to more than 400 instrument manufacturers, are

1. Uniformity of tube characteristics.
2. Tube life.
3. Increased operation stability.
4. Reduction of microphonic effects.
5. Reduction of hum.
6. Lower noise level
7. Type of tube envelope.
8. Service conditions.
9. Yearly demand.
10. Cost of tubes.

The results of this survey will be tabulated and published as a committee report, and readers of *Electrical Engineering* are requested to assist in making it a comprehensive survey. Companies which have not received questionnaires may request them from the AIEE Joint Subcommittee on Electronic Instruments, AIEE Headquarters, 33 West 39th Street, New York 18, N. Y.

DISTRICT • • • • •

Southern District Students Convene in 1947

Tentative plans have been announced for a convention of the Student Branches of the Southern District to be held about the first week end in April 1947. The convention will open Thursday morning and extend through Saturday.

Convention headquarters will be on the campus of the University of Kentucky, Lexington. A number of student papers are expected to be submitted for prizes, and it is hoped that every Student Branch will be represented by at least one paper.

Included in the scheduled program are several speakers, a Thursday evening social, the reading of the student papers, tours of the bluegrass region, and a banquet and dance on Friday night. Special entertainment is planned for the women guests.

SECTION • • • • • •

Wire Strain Gauges Discussed by Canton Subsection

Professor K. F. Sibila (A '38) of the University of Akron (Ohio) was speaker at the first technical meeting of the Canton Subsection held October 14, on the premises of the Ohio Power Company.

Drawing on his experiences with the Guggenheim Institute during the war, Professor Sibila discussed the application of wire strain gauges to models in wind tunnels and the use of hot wires in the measurement of air velocities and directions.

ABSTRACTS • • • •

TECHNICAL PAPERS previewed in this section will be presented at the AIEE winter meeting, New York, N. Y., January 27–31, 1947, and will be distributed in advance pamphlet form as soon as they become available. Members may obtain copies by mail from the AIEE order department, 33 West 39th Street, New York 18, N. Y., at prices indicated with the abstract; or at five cents less per copy if purchased at AIEE headquarters or at the meeting registration desk. Prices for nonmembers will be twice those for members.

**Mail orders will be filled
AS PAMPHLETS BECOME AVAILABLE**

Air Transportation

47-5—Parallel Operation of Aircraft Alternators Using Electronic Frequency Changers; *O. W. Bowlus (M '46), P. T. Nims (A '40). 25 cents.* This paper reports initial work on an electronic method for operating several main aircraft engine

Attending the banquet held September 21, 1946, to celebrate the opening of the new headquarters of the Engineers' Society of Milwaukee were AIEE President J. Elmer Housley (F '43) and the following members of the Milwaukee Section: (front row, left to right) L. M. Swenson (A '41) Allis-Chalmers Manufacturing Company, chairman of the Section program committee; F. J. Van Zeeland (M '44) professor at the Milwaukee School of Engineering, and secretary–treasurer of the Section; Mr. Housley; K. L. Hansen (F '34) Harnischfeger Corporation, past chairman of the Section; E. L. McClure (M '41) Wisconsin Electric Power Company, director of the Section; L. C. Aicher, Jr. (M '43) Allis-Chalmers Manufacturing Company, director of the Section. In the second row are: C. P. Feldhausen (M '40) Cutler-Hammer, Inc., director of the Section; S. H. Mortensen (F '20) Allis-Chalmers Manufacturing Company, past Section chairman; W. C. Johnson, Allis-Chalmers Manufacturing Company; Walther Richter (F '42) Allis-Chalmers Manufacturing Company, Section chairman; E. J. Limpel (M '44) A. O. Smith Corporation, director of the Section; E. T. Sherwood (M '44) Globe-Union, Inc., director of the Section; H. C. Brem (A '43) Allis-Chalmers Manufacturing Company, director of the Section

driven alternators in parallel to supply the electrical loads of an airplane. This would make unnecessary any mechanical constant speed drive for each generator. An electronic frequency changer is associated with each alternator and the outputs of the changers paralleled at the load bus. Synchronization, load division, and provision for reactive power are discussed, and test data obtained from two 4-kva units are presented.

47-11—The Calibration of Ignition Crest Voltmeters; *W. L. Davis, C. E. Warren. 15 cents.* Field experience with crest voltmeters for use on internal combustion engine ignition systems has shown variation in readings between several types of meters connected to the same point in an ignition system. This paper outlines a study of the factors influencing the performance of several types of meters. Apparatus is described for producing repetitive voltage surges the time of which to crest, shape of tail, amplitude, and repetition rate are easily adjustable over a considerable range. The equipment was used to determine the effects of the variables mentioned on the operation of several types of ignition crest voltmeters. A high-speed cathode-ray oscillograph was used to calibrate the meters on various types of voltage surges. Typical results for two meters are given in the form of calibration areas. Effects of voltage wave shape on the meter readings are discussed.

Basic Sciences

47-1—Eddy Currents in Disks: Driving and Damping Forces and Torques; *A. D. Moore (F '43). 30 cents.* Driving torque due to a-c quadrature fluxes through a disk can be solved for by means of theory and procedures presented. Fluxes may have any distribution. A facilitating chart is described. As an example, the driving torque of a watt-hour meter is solved for. Damping torque of a disk rotating through constant fluxes can be solved for by means of theory and procedures developed from what may be a new concept. Flux areas may be irregular. The methods can be extended to cover cases of nonuniform density. A facilitating chart is described. As an example, the damping torque of a watt-hour meter is solved for. Damping of disks moved translationally is covered by a family of curves, by which damping for accelerometers and other devices can be designed in direct fashion. For both driving and damping torques, reciprocity laws are proved.

47-2—Theory and Application of Complex Logarithmic and Geometric Mean Distances; *T. J. Higgins (M '46). 20 cents.* Logarithmic and geometric mean distance theory is utilized both in the design and in the calculation of the resulting electrical and mechanical performance of power and transmission lines, industrial distribution busses, transformer windings, coils and solenoids for various purposes, and yet other current-carrying apparatus. Heretofore calculation of the required logarithmic and geometric mean distances has been effected by direct integration of the multiple integral,

$$\int_{S_2}\int_{S_1} \log\,[(X-x)^2+(Y-y)^2]^{1/2}\,dS_2\,dS_1$$

wherein (X, Y) and (x, y) indicate arbitrary points in the lineal or areal cross sections S_1 and S_2 of the current-carrying apparatus. Except in certain simple well-known cases, evaluation of this multiple integral requires very heavy analysis. This difficulty is obviated by use of the theory advanced in this paper. By use of simple conformal mapping, the foregoing real integral is expressed as the real or imaginary part of a complex integral which easily is evaluated. The resulting complex logarithmic mean distance comprises as its readily recognized real or imaginary part the desired real logarithmic and geometric mean distances. Calculation of several logarithmic mean distances typical of those used in practice illustrates application of the general theory and the advantages to be gained by use of it.

Electric Machinery

47-3—Equivalent Circuit of the Primitive Rotating Machine With Asymmetrical Stator and Rotor; *Gabriel Kron (M '45). 20 cents.* The equations and equivalent circuits of all standard types of rotating electric machinery operating under arbitrary conditions may be derived as special cases of a single generalized rotating machine, the so-called primitive machine. This paper derives the steady-state equivalent circuit of the primitive machine having *unbalanced rotor* as well as stator. That is, the load or transmission line connected to a salient-pole synchronous machine is assumed to be unbalanced (due to faults, short circuits, and the like) or the load connected to the slip rings of an induction motor is assumed to be unbalanced. The single-phase alternator, instrument Selsyns and so forth, as well as sudden short-circuit studies are cases that require an infinite series of time harmonics in the stator and rotor windings in addition to the fundamental. It is assumed that these time harmonics are due to space waves with fundamental pairs of poles rotating at different speeds. The effect of the harmonic space waves is not considered. A companion paper by Rankin shows network analyzer studies made of the harmonic currents and torques of a single-phase alternator.

47-9—The Dielectric Properties of Cellulose Insulation Impregnated With Semiconductor Liquids; *F. M. Clark (A '24). 20 cents.* Liquid impregnants for cellulosic insulation are classed as dielectric, semiconducting, and conducting. The low power factor of insulation impregnated with a dielectric liquid rises to high value as the liquid resistivity decreases from the dielectric to the semiconducting classification. In the semiconducting grouping, there is a limited range of resistivity which is accompanied by a marked drop in the power factor of the treated insulation. With kraft paper, the power factor falls to about three per cent. Because of the conducting nature of the liquid, the effective dielectric constant obtained simultaneously with such low power factor is abnormally high. The Permalytic capacitor is a kraft paper spaced capacitor of the usual type, impregnated with a semiconducting liquid. This capacitor is characterized by a physical size equal to about 30 per cent of the physical volume per kilovolt-ampere of the corresponding oil-treated capacitor. The power factor of the Permalytic capacitor is about three per cent. This type of capacitor offers interesting possibilities for those applications where a small dependable low-cost low-voltage capacitor is required.

47-10—Simplified Graphical Method of Computing Thermal Transients; *Paul Narbutovskih (M '43). 15 cents.* During recent years the subject of estimating the transient overload capacity of electric equipment, transformers in particular, has gained considerable importance among operators and designers. Practically acceptable methods of making such estimates have been developed, including graphs and charts, which simplify the computations of certain specific problems involved. This paper gives a new graphical method of computing thermal transients which possesses a considerable degree of simplicity and flexibility. The method is based on the use of new, here proposed, transient co-ordinate paper on which any single-capacity thermal or electrical exponential transient will plot as a straight line. The use of this paper greatly simplifies the solution of an exponential equation for temperature and time. It adapts itself readily to handling transient temperatures in case of complex load cycles, as well as more intricate problems involved in equipment design.

Instruments and Measurements

47-12—A Comparison of Two Basic Servomechanism Types; *Herbert Harris (A '46). 30 cents.* Servomechanisms with lead network stabilization and those with speed feedback through a high pass filter are analyzed and compared. The method of analysis is on a frequency response basis using inverse transfer functions. Detailed design tabulations are given along with the limitations imposed by response delays. Analysis of motor suitability for these servomechanisms shows that torque-to-inertia ratio is a poor quantity for a figure of merit, and several alternative ratios are shown to have more value.

47-7—The Application of Lead Networks and Sinusoidal Analysis to Automatic Control Systems; *G. J. Schwartz (A '46). 30 cents.* This paper discusses the general design considerations relating to the construction of a high performance servomechanism. The sinusoidal method of analysis is applied, and transfer loci for complete servo systems and their individual components are presented in order to illustrate the theoretical results obtained. Particular emphasis is placed upon the effect of anticipation or lead networks on the performance of a servomechanism. Direct reference to an existing armature-controlled d-c motor servomechanism is made throughout, and the correlation between theory and practice as applied to a servomechanism design is demonstrated.

47-13—The Cathode-Ray Spectrograph; *Rudolf Feldt (M '46), Carl Berkley. 15 cents.* The instrument described has been developed by the Allen B. DuMont Laboratories Inc., and produces complete spectral distribution curves of the source under test at the rate of 240 complete spectra per second. Spectrogram is presented on a cathode-ray tube. The spectrograph consists of four units which may be used independently. These are: source, capacitating system, scanning means, and indicator. Wave length calibration is done by replacing source with a gaseous discharge lamp. Sensitivity is adjusted so that resultant curve represents visual response. Dispersion is given by prism with scanning effected by a synchronous, vibrating, resonant mirror at 120 cycles per second. A suitably filtered multiplier photoelectric cell is used. A standard cathode-ray oscillograph may be used as indicator, but special d-c oscillographs will be available for this purpose. Samples may be compared by switching or use of delay screens. This instrument was developed initially to study the color of microsecond current pulses on a cathode-ray tube fluorescent screen. Results are shown. The color of any other varying light source can be studied similarly. Provision is made for insertion of transmission samples, chemical samples, or filters for calibration. Continuous recording is possible with camera attachment. Indication may be remote or separated from analyzer. Possible applications are:

1. Instantaneous color matching.
2. Fluorescent lamp characteristics.
3. Spectrographic research.
4. Continuous process control.
5. Transmission of spectral characteristics from remote or inaccessible locations.

Power Generation

47-4—Theoretical Approach to Speed and Tie Line Control; *Robert Brandt (A '26). 15 cents.* Frequency error is an indicator of the amount of excess or deficiency of generation on a whole interconnection, and a frequency-generation line may be established for the whole or for any section which is separated from all others by tie lines over which control may be desired. Automatic equipment is available, using these fundamentals under the name of load biased frequency control, which will make the required adjustments to generation to restore normal speed and tie line loadings with a minimum of false moves. By this means each section will handle, in normal operation, only its own load changes, regardless of the speed of response of its prime movers to correcting impulses. Methods of automatic control by the selective-frequency blocking system and the master flat-frequency station method are examined and found wanting. The use of uncontrolled speed-sensitive governors with small droop is shown to be disadvantageous, if operated in parallel with automatic bias equipment. Consideration is given to practical matters of operation, including an improvement in handling manual generation shifts, action of controller in time of trouble, effect of errors in determining the frequency-generation line, and desirable characteristics of governors.

Land Transportation

47-8—Developments in Current Collectors for High Speed Service; *B. F. Langer. 15 cents.* Theoretical studies and wind-tunnel tests have been used to develop an improved pantograph shoe and shoe mounting. The shoe itself has been made aerodynamically stable, and the supporting springs make it follow the wire smoothly at all speeds.

Protective Devices

47-6—Resistors for 138-Kv Cable Switching; *E. K. Sadler (M '29), T. M. Blakeslee (M '38). 25 cents.* Installations of 138-kv underground cable by the Los Angeles Department of Water and Power provided a field of laboratory for capacitive-current switching tests. This paper describes and presents results of an investigation of overvoltages incident to switching the charging current of these cable circuits. The development and performance of step-type resistors installed in 138-kv circuit breakers for limiting these overvoltages are described. A theoretical discussion of the effect of resistors in circuit breakers handling capacitive currents is given in an appendix.

PERSONAL • • • • •

L. W. Chubb

Lewis Warrington Chubb (A '09, F '21) director of research, research laboratories, Westinghouse Electric Corporation, East Pittsburgh, Pa., has been awarded the John Fritz Medal for 1947 for "pioneering genius and notable achievements during a long career devoted to the scientific advancement of the production and utilization of electric energy." Doctor Chubb was born October 22, 1882, in Fort Yates, N. Dak., and was graduated from Ohio State University in 1905 with the degree of mechanical engineer in electrical engineering. He became an engineer apprentice with the Westinghouse corporation and from 1906 to 1910 worked on the development and testing of sheet steel and did research on transformers. He was placed in charge of the electrotechnical and magnetic research section in 1910 and of the research work of the material and process engineering department in 1919. When the radio engineering department was organized in 1920, Doctor Chubb was made manager. He left East Pittsburgh in 1930 when the radio activities of the Westinghouse corporation, the General Electric Company, and the Radio Corporation of America were consolidated at Camden, N. J., to become assistant to the vice-president of RCA. Later that year he returned to East Pittsburgh as director of research. Doctor Chubb's contribution to the knowledge of magnetic properties of iron and iron alloys, and his improvements in the design of electric machinery and in the measurements of electrical and magnetic quantities have had great influence in the development of electrical engineering during the past 30 years. He has been granted about 200 patents for electrical, mechanical, chemical, electrochemical and instrumental developments. He has received the degree of doctor of science from Allegheny College and from the University of Pittsburgh. In 1934 he was awarded the Lamme Medal of Ohio State University. Author of many technical articles, he is a member of the American Association for the Advancement of Science, the American Physical Society, the Franklin Institute, the Institute of Radio Engineers, the Illuminating Engineering Society, and the National Electrical Manufacturers Association. He also is a member of Tau Beta Pi and Sigma Xi. He formerly was a member of the International Electrotechnical Commission.

C. A. Robinson (A '11, F '22) vice-president, secretary, and treasurer, Chesapeake and Potomac Telephone Company, Washing, D. C., has retired. Mr. Robinson received the degree of bachelor of arts from Johns Hopkins University in 1903 and the degree of mechanical engineer from Cornell University in 1906. Immediately afterwards he entered the engineering department of the New York Telephone Company, New York. From 1909 to 1919 he was employed in the telephone transmission branch of the engineering department of the American Telephone and Telegraph Company, New York. In 1919 he was sent by the Western Electric Company to Europe in connection with engineering a large toll telephone cable project in France. He was appointed transmission engineer of the American Telephone company later that year and in 1920 became chief engineer of the Chesapeake company. He was appointed assistant vice-president in 1929 and general manager of the West Virginia company in 1931. He was elected vice-president and general manager of the Washington company in 1940.

L. A. Kilgore (A '29, M '37) formerly head of the rectifier and motor section of the a-c generator department, Westinghouse Electric Corporation, East Pittsburgh, Pa., has been appointed assistant manager of the a-c generator department. Mr. Kilgore joined the corporation in 1927 after his graduation from the University of Nebraska. Mr. Kilgore is an ex-chairman of the AIEE committee on electric machinery and a member of Sigma Xi. **M. B. Lory** (A '40) has been named manager of the motor section of the department and **J. L. Boyer** (A '43) manager of rectifier development. Mr. Lory, who holds the degree of bachelor of science from Colorado State College and the degree of master of science from Massachusetts Institute of Technology, joined the Westinghouse corporation in 1928. Mr. Boyer has been with the corporation since he was graduated from Rice Institute in 1942. He is a member of Tau Beta Pi.

L. R. Milburn (A '20, F '39) electrical engineer, Great Lakes Steel Corporation, Ecorse, Mich., has been elected president of the Association of Iron and Steel Engineers. Mr. Milburn was graduated from the University of Michigan in 1918, and, after service in the United States Army, he entered the test course of the General Electric Company at Schenectady, N. Y., and Pittsfield, Mass. He was transferred to the Cleveland, Ohio, office in 1919 as complaint and installation engineer and to the Cincinnati, Ohio, office in 1923. From 1925 to 1929 he was engaged in construction work for the Louisville Gas and Electric Company. He joined the Great Lakes company as electrical draftsman in 1929, became acting electrical engineer in 1933 and electrical engineer in 1935.

P. H. Maurer (A '33) formerly commercial application engineer, National Pneumatic Company, New York, N. Y., has been appointed chief engineer of the company. Mr. Maurer attended Newark College of Engineering and the University of Michigan. He previously was associated with the Bendix Aviation Corporation and with the Hudson Motor Car Company as electrical engineer in charge of car engineering. At the outbreak of the war he was loaned to Sparks-Withington Company, Jackson, Mich., which was building link trainers but later he returned to the Hudson company which was operating the Naval Ordnance Plant, Centerline, Mich. He was appointed executive engineer of the Redmond Company, Owosso, Mich., and in 1945 joined the National Pneumatic Company. He is a member of the Engineering Society of Detroit and the Society of Automotive Engineers.

E. T. Augustine (A '27, M '33) formerly assistant chief engineer, Western Massachusetts Electric Company, Springfield, has been named assistant general superintendent of the company. Mr. Augustine was employed by the company in 1929 as system planning engineer. He was district engineer from 1932 to 1944 in which year he was appointed assistant chief engineer. **M. G. Moses** (A '41, M '42) formerly planning engineer, has been appointed assistant chief engineer of the system. A 1927 graduate of the University of Minnesota, he was employed in the radio research laboratory at Harvard University, Cambridge, Mass., from 1943 to 1945, the year he joined the Western Massachusetts company. Previously he had been associated with the Northwestern Public Service Company and the Northern States Power Company.

A. B. Campbell (A '20, F '38) formerly eastern representative of Hughes Brothers, New York, N. Y., has been appointed executive secretary of the National Association of Corrosion Engineers with headquarters in Houston, Tex. Mr. Campbell holds the degrees of bachelor of science and electrical engineer from the University of Illinois. He joined the faculty of Iowa State College in 1915 and was appointed to the Iowa Board of Railroad Commissioners in 1920. He became associated with the National Electric Light Association in 1924 and continued with its successor, the Edison Electric Institute, until 1943, at which time he joined Hughes Brothers.

D. C. Inman (A '38, M '45) formerly engineering and service manager for the Westinghouse Electric Corporation, Cleveland, Ohio, has been appointed engineering manager of the district engineering and service department, East Pittsburgh, Pa. A graduate of Colorado Agricultural College, Mr. Inman joined the company in 1925. After ten years in the control engineering department, he joined the Pittsburgh, Pa., engineering division of the company as consulting and application engineer. In 1938 he was transferred to the Cleveland office and became manager in 1941. He is a past president of the Cleveland Engineering Society.

Hector Sleeman (M '27, F '36) colonel in the British Army, who during the war served in Burma and India with the Royal Artillery and the Indian Electrical and Mechanical Engineers, will be released from active service in January 1947. After a trip to engineering establishments in Great Britain, Canada, and the United States, Colonel Sleeman intends opening a consulting firm in Melbourne, Australia. Before entering military service in 1939, Colonel Sleeman was chief engineer of the Rangoon (Burma) Electric Tramways and Supply Company.

J. O. Johnson (A '36, M '43) formerly chief product engineer, product engineering division, Aircraft-Marine Products, Inc., Harrisburg, Pa., has been appointed assistant general manager of the Buchanan Electrical Products Company, Inc. Mr. Johnson, who was graduated from Rensselaer Polytechnic Institute in 1931, was associated with Sears Roebuck and Company, Chicago, Ill., until 1933. He was assistant to the chief electrical engineer at the Material Laboratory, United States Navy Department, Brooklyn, N. Y., from 1934 to 1940, and became chief product engineer with Aircraft-Marine Products in 1941.

W. D. Coolidge (A '10, M '34) retired director of the General Electric Research Laboratory, Schenectady, N. Y., has been appointed to head the new General Electric atomic research and development laboratory at the Hanford engineer works near Richland, Wash. Doctor Coolidge has been X-ray consultant to the General Electric Company since his retirement in 1945 (*EE, Mar '45, p 125*). He has been associated with the atomic bomb project since 1941 at which time he was one of the six men selected to evaluate the military importance of uranium.

W. A. MacCrehan, Jr. (A '43) formerly supervisor of quality control in the motor division, General Electric Company, Lynn, Mass., has been appointed assistant professor of administrative engineering and director of the Gage Laboratory at New York University, New York, N. Y. Mr. MacCrehan has been instructor for the State University extension courses in quality control at Harvard University, Cambridge, for the past three years and at Boston University Evening College of Commerce for 1945–46. He is a member of the National Society for Quality Control, the American Standards Association, and the American Society of Tool Engineers.

S. G. Lutz (A '38) formerly head of the measurements and direction-finding section of the Naval Research Laboratory, Washington, D. C., has been appointed chairman of the electrical engineering department of the College of Engineering of New York (N. Y.) University. Doctor Lutz received from Purdue University the degree of bachelor of science in 1933, master of science in 1934 and doctor of philosophy in 1938. He was assistant professor at Southern Methodist University, Dallas, Tex., from 1938 to 1940. In 1940 he joined the Naval Laboratory. He is a member of the Institute of Radio Engineers and the American Society for Engineering Education.

G. L. Rosenberger (A ’31) superintendent of construction, Virginia Electric and Power Company, Alexandria, has retired. Mr. Rosenberger has been with the Virginia Power company and its predecessors since 1924. Previously he was superintendent of public utilities for Manassas, Va., from 1914 to 1920 and superintendent of the Herndon Light and Supply Company from 1920 to 1922. As superintendent of the Fairfax-Loudon Light and Power Company from 1922 to 1924, he built the first transmission line between Alexandria and Leesburg, Va.

W. G. Worcester (A ’42) engineer, general engineering laboratory, General Electric Company, Schenectady, N. Y., has been appointed assistant professor of electrical engineering at the University of Colorado, Boulder. Professor Worcester was graduated from the University of Colorado in 1939 with the degree of bachelor of science and received the degree of master of science from California Institute of Technology in 1940. He has been with the General Electric Company since 1941. He is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

E. R. McKee (A ’30, M ’36) professor and head of the department of electrical engineering, University of Vermont, Burlington, and acting dean of the college of engineering, has been named dean of the new College of Technology at the University. Professor McKee, who holds the degrees of bachelor of science, master of science, and electrical engineer from Iowa State College, has been a member of the faculty of the University of Vermont since 1934.

T. J. Bostwick (A ’18, M ’26) chief electrical engineer, Aluminum Company of America, Pittsburgh, Pa., has retired. However Mr. Bostwick will remain available to the company and its subsidiaries as consulting engineer. Mr. Bostwick was born in 1876 in Oswego, N. Y. After two years in the students’ course at the Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa., and three as electrician for the Niagara Falls (N. Y.) Power Company, he joined the Aluminum Company in 1900 as electrical engineer. In 1912 he was made assistant chief electrical engineer and in 1924 chief electrical engineer.

H. E. Kent (A ’25, M ’31) engineer with the Edison Electric Institute, New York, N. Y., has been appointed director of engineering. Mr. Kent, who was born in 1901 in Providence, R. I., holds the degrees of bachelor and master of science from Massachusetts Institute of Technology. Mr. Kent has been associated first with the National Electric Light Association and then with the Edison Electric Institute since 1924. He has been concerned particularly with the problem of the inductive co-ordination of power and communication circuits. He is a member of Tau Beta Pi.

F. E. Nutt (A ’36) formerly assistant sales manager, Crocker-Wheeler Electric Manufacturing Company, division of Joshua Hendy Iron Works, Ampere, N. J., has been appointed executive engineer of the division. Mr. Nutt is a 1931 graduate of the University of Denver. Before joining the Crocker-Wheeler sales department in 1942 he was a student engineer and assistant in the high voltage laboratory of the General Electric Company, Pittsfield, Mass. He was Crocker-Wheeler sales representative in Washington, D. C., from 1943 until in 1945 he was made assistant sales manager.

C. S. Allen (A ’31, M ’37) formerly assistant general manager, Sawyer Electrical Manufacturing Company of Los Angeles, Calif., has been elected vice-president and general manager of the company. Mr. Allen was graduated from Mississippi State College in 1929 and, after completing the test course at the General Electric Company, was assigned to the fractional horsepower motor engineering department as design engineer. He joined the Sawyer Company as chief engineer in 1945.

R. C. Mason (A ’26, M ’37) of the electrophysics department, Westinghouse Research Laboratories, East Pittsburgh, Pa.; **J. W. Simpson** (A ’43) section engineer of the company’s switchboard engineering department; and **A. H. Toepfer** (A ’31, M ’39) application engineer in the district engineering and service department; have been loaned for one year to the Institute of Nuclear Studies which is being established at Oak Ridge, Tenn., to provide channels for co-operative research between government, universities, and industrial concerns.

E. I. Smith (A ’32) until recently with the Signal Corps, has been appointed telegraph superintendent of the Atlantic Coast Line Railroad Company, New York, N. Y. Mr. Smith was with the Western Union Telegraph Company from 1931 until he entered military service in 1942. During the war he served as chief signal officer, radio station WAR, Washington, D. C., and spent 34 months in Africa and Europe.

R. W. Harper (A ’26, M ’36) of Bell Telephone Laboratories, Inc., New York, N. Y., and recently lieutenant colonel and commanding officer of the 71st Region, Plant and Engineering Agency, Army Communications Service, has received the Army Commendation Ribbon. The citation states that he “capably discharged important responsibilities in directing and co-ordinating the construction and installation of radio communication facilities and navigational aids for aircraft to complete communications along the routes of the Air Transport Command.”

A. W. Post (A ’06, M ’28) equipment cost studies engineer, long lines department, American Telephone and Telegraph Company, New York, N. Y., has retired. Mr. Post was graduated from the Polytechnic Institute of Brooklyn in 1909 and that year entered the plant department of the long lines department. He was transferred to the engineering department in 1917. In recent years he has been concerned with costs of central office and other equipment associated with the long lines plant.

H. C. Purcell (A ’41) formerly electrical designer, engineering department, Hawaiian Howard Electric Company, Honolulu, T. H., and **C. R. Day** (A ’44) formerly assistant engineer, Modesto (Calif.) Irrigation District, have been appointed assistant designing engineers for the Sacramento (Calif.) Municipal Utility District. Mr. Purcell is a graduate of Stanford University and Mr. Day of the University of California.

D. H. Lauder (M ’45) formerly assistant manager, service engineering division, General Electric Company, Schenectady, N. Y., has been appointed works manager of the company’s newly acquired Hanford (Wash.) engineer works (*EE, July ’46, p 361*). The atomic energy research in progress at Hanford has been designated the General Electric Nucleonics Project. Mr. Lauder has been with the company since he was graduated from the University of Pittsburgh in 1922.

A. B. Morgan (A ’28, M ’35) formerly rate and power consultant, Edison Electric Institute, New York, N. Y., has been appointed managing director of the institute. A graduate of Massachusetts Institute of Technology, Mr. Morgan joined the National Electric Light Association, predecessor of EEI, in 1927 as an assistant engineer. He was titled engineer and rate and power consultant in 1933.

A. W. Melloh (A ’38, M ’45) senior telephone engineer, Stromberg-Carlson Company, Rochester, N. Y., has been appointed to the telephone transmission group and will be responsible for the development of carrier-current equipment. Mr. Melloh, a graduate of the University of Minnesota, trained officers and enlisted personnel in the operation and maintenance of underwater sound equipment at the Navy’s San Diego (Calif.) laboratory during the war.

J. W. Mullally (A ’45) formerly manager of bare and weatherproof sales for the Anaconda Wire and Cable Company, New York, N. Y., has been appointed manager of utility sales. Previously Mr. Mullally served as district manager for the company in Pittsburgh, Pa. During the war he served for three years as chief of the Wire Mill Branch of the Copper Division of the War Production Board.

D. I. Bohn (A ’23, M ’41) formerly electrical engineer, has been appointed chief electrical engineer of the Aluminum Company of America, Pittsburgh, Pa. Mr. Bohn, who was graduated from the University of Wisconsin in 1921, entered the employ of the company as assistant electrical superintendent in Badin, N. C., in

1923 and was made superintendent in 1925. He was transferred to Pittsburgh as electrical engineer in 1928.

E. A. Ossmann (A '44) electronic engineer, Allen B. Du Mont Laboratories, Inc., has been named New York State factory representative for the company. Mr. Ossmann will have his headquarters in Rochester, N. Y. A graduate of Manhattan College, Mr. Ossmann previously was associated with the Anaconda Copper Corporation and the General Electric Company and served in the Armed Forces.

G. P. Lehmann (A '35) formerly colonel, 12th Air Force Service Command, has been appointed assistant manager of the plastics division, General Electric Company, Pittsfield, Mass. Mr. Lehmann had been with the General Electric Company from 1935 until he went on active duty as a lieutenant in 1941.

J. H. Howard (A '41) research supervisor, Engineering Research Associates, St. Paul, Minn., has received the award of the Legion of Merit from the United States Navy in recognition of outstanding service performed while on duty with the Division of Naval Communications. Mr. Howard served in the Navy from 1943 to 1946 and was a lieutenant at the time of his release from duty.

F. B. Silsbee (A '13, F '42) formerly chief of the electric instruments section, National Bureau of Standards, Washington, D. C., has been appointed chief of the division of electricity of the bureau. Doctor Silsbee, who holds the degrees of bachelor and master of science from Massachusetts Institute of Technology and received the degree of doctor of philosophy from Harvard University in 1915, has been with the National Bureau of Standards since 1911. He became chief of the electric instruments section in 1939.

W. R. Lyon (A '20, M '26) recently lieutenant colonel in the Army Air Forces, Wright Field, Ohio, has returned to Bell Telephones Laboratories, Inc., New York, N. Y. For about half of his five and a half years in the Army, he was resident representative in a number of manufacturing plants in Ohio and Indiana. For over two years he was successively instructor in electrical engineering at the AAF Engineering School and professor at the AAF Institute of Technology, Wright Field.

D. E. B. Corson (A '33, M '41) formerly sales manager of the power division of Aerovox Corporation, New Bedford, Mass., has been appointed manager of the special products division of Solar Manufacturing Corporation, New York, N. Y. A 1932 graduate of Massachusetts Institute of Technology, Mr. Corson previously was sales manager of the power factor division of Cornell-Dublier Electric Corporation, South Plainfield, N. J.

Mary F. Blade (A '35) who has been appointed instructor in machine drawing at Cooper Union, New York, N. Y., will be the first woman faculty member in the school of engineering in 87 years. Mrs. Blade, who was graduated from the University of Utah in 1934, since has been electrical engineer for the Southern Utah Power Company, Cedar City, and methods engineer for the United States Employment Service.

L. W. Whitton (A '40) formerly production manager, Otis Elevator Company, New York, N. Y., has been appointed manager of operations. A 1922 graduate of the University of California, Mr. Whitton joined the company in 1919 in San Francisco, Calif., was transferred to New York in 1936, and has been production manager since 1944.

W. E. Tinsley (A '44) works engineer, Sperry Gyroscope Company, Great Neck, N. Y., is one of the first three engineers to receive graduate scholarships under a newly instituted plan of the company. Mr. Tinsley is studying for the degree of master of science in aeronautical engineering at New York University.

D. A. Quarles (A '23, F '41) director of apparatus development, Bell Telephone Laboratories, Inc., New York, N. Y., has become a member of the committee on electronics of the Joint Research Board. Three officers each from the Army and Navy and two other civilians complete the committee.

E. E. Browning, Jr. (M '38) formerly plant extension engineer, operations and engineering department, American Telephone and Telegraph Company, New York, N. Y., has been placed in charge of a new section on engineering results in that department with the title of results engineer.

G. W. Morgan (A '37) formerly design electrical engineer, Gulf States Utilities Company, Beaumont, Tex., has joined the Hawaiian Electric Company, Ltd., Honolulu. Mr. Morgan, who joined the Gulf States company in 1940, previously had worked for Stone and Webster Engineering Corporation.

E. M. Strong (A '26, M '40) professor and chairman of the electrical engineering division, Cornell University, Ithaca, N. Y., has been appointed to represent the Illuminating Engineering Society in the division of engineering and industrial research of the National Research Council.

Edward Kerschner (A '19) formerly eastern district manager, General Cable Corporation, New York, N. Y., is now president of The Electrical Distributors Company, Philadelphia, Pa., which represents electrical manufacturers in the Philadelphia, Pa.; Baltimore, Md.; and Washington, D. C., areas.

OBITUARY

Henry Metcalf Hobart (A '94, M '99, F '12) consulting engineer, Schenectady, N. Y., died October 11, 1946. Born November 29, 1868, in Boston, Mass., Mr. Hobart was graduated from Massachusetts Institute of Technology with the degree of bachelor of science in electrical engineering in 1889. That same year Mr. Hobart commenced his career with the Thomson-Houston Electric Company, Lynn, Mass. In 1894 he went to Schenectady, N. Y., with the General Electric Company, where he was assistant to C. P. Steinmetz, who, in recommending Mr. Hobart for AIEE membership, called him "one of the best electrical engineers I ever met." He returned to Lynn in 1895, remaining until in 1896 he joined the British Thomson-Houston Company, London, England. In 1900 he became associated with the Union Electricitats Gesellschaft in Berlin, Germany. From 1903 to 1911 he conducted a private consulting practice in London, afterwards returning as consulting engineer to the General Electric Company. He retired in 1940. Mr. Hobart was a lecturer at the University College, London, from 1908 to 1911 and lectured also at Northampton Institute and Faraday House in London, and he was external examiner for the University of Sheffield for a number of years. He delivered the James Forrest Lecture before the Institution of Civil Engineers in 1915. He was author and coauthor of more than a dozen books, some of which were translated into French and German, and was also author of many technical papers. At least 30 patents were issued in his name. In 1936 he received the Samuel Wylie Miller Memorial Medal of the American Welding Society. Mr. Hobart maintained a life-long interest in AIEE activities. He served as manager for 1922–26, vice-president for 1926–28, and chairman of the Schenectady Section for 1914–15. At various times he was member of the following AIEE committees: executive, Edison Medal, Standards, electric machinery, electric welding, research, and transportation. He represented the AIEE on the American Bureau of Welding, the American Engineering Council, the American Standards Association, and the United States National Committee of the International Electrotechnical Commission. He was a member of the American Welding Society, the American Society of Mechanical Engineers, the American Association for the Advancement of Science, the Institution of Civil Engineers, the Institution of Mechanical Engineers, and the Institution of Electrical Engineers.

Sylvester Bedell Way (A '03, F '38) chairman of the boards of the Wisconsin Electric Power Company, the Milwaukee Electric Railway and Transport Company, the Wisconsin Gas and Electric Company, and the Wisconsin Michigan Power Company, Milwaukee, Wis., died September 20, 1946. Born August 29, 1874, in Philadelphia, Pa., Mr. Way was graduated from Drexel Institute in 1896. He was awarded the honorary degree of doctor of science by that institution in 1942. After a brief period with the Electric Storage Battery Company, Philadelphia, he became chief electrical engineer and superintendent of what is now the Union Electric Light and Power Company, St. Louis, Mo., in 1898. In 1911 he was appointed assistant general manager of the Milwaukee Electric Railway and Light Company. He became vice-president and general manager in 1914 and president in 1925. He relinquished the position of general manager in 1934 and was elected chairman of the board in 1945. It was under Mr. Way's direction that the company's Lakeside power plant was designed and became the first large central station constructed for the exclusive use of pulverized fuel. In 1917 he was given direct supervision of the company's street railway, and to solve the man power problem during World War I, he designed and patented the center truck construction for two-car articulated trains. Mr. Way was director of the First Wisconsin National Bank, the First Wisconsin Trust Company, the Milwaukee Association, the the Milwaukee Community Fund, the Council of Social Agencies, and the Milwaukee Foundation. He was a member of the Engineers' Society of Milwaukee.

William H. Matthies (M '26, F '29) retired systems engineer for Bell Telephone Laboratories, Inc., New York, N. Y., died October 20, 1946, in Hackensack, N. J. Mr. Matthies was born January 28, 1879, in Hamburg, Germany, and was graduated from Massachusetts Institute of Technology in 1902 with the degree of bachelor of science in electrical engineering. After graduation he joined the Western Electric Company as student engineer and was titled equipment engineer in 1903 and as such began to specialize on circuit design work. At that time he worked on the first dial exchange in the United States. He was appointed chief engineer of the company's Berlin, Germany, branch in 1905 and while in that position engineered and placed in service the first common battery switchboards in use in Germany and Scandinavia. He was attached to the staff of the European chief engineer of the company from 1911 to 1916 at Antwerp, Belgium, specializing in machine switching development. He returned to New York in 1916 and was placed in charge of machine switching circuit development. Under his leadership both the crossbar system of dial exchanges and the earlier panel system were developed. In 1921 he was given charge of all local central office systems development and of private branch exchange development and after 1925 continued with that work for Bell Telephone Laboratories. Mr. Matthies held more than 30 patents now in use in Bell equipment. He retired in 1943.

John Winchell Creasey (A '20, M '29) division plant superintendent, plant department, American Telephone and Telegraph Company, St. Louis, Mo., died recently. Born January 10, 1892, in Rich Hill, Mo., Mr. Creasey received the degree of bachelor of science in electrical engineering from the University of Missouri in 1914. After a year as student apprentice with the General Electric Company, Fort Wayne, Ind., Mr. Creasey was employed briefly as electrician for the Chicago, Burlington, and Quincy Railway, North Kansas City, Mo., and by Peet Brothers Manufacturing Company, Kansas City, Kans. He entered the Bell System in 1916 as equipment attendant for the American Telephone and Telegraph Company, Kansas City, Mo., and the following year was transferred to St. Louis as inspector. In 1919 he was made district plant manager in Kansas City and from 1922 to 1924 he was assigned to the general plant manager's office in New York, N. Y. He returned to Kansas City as district plant superintendent in 1924 and in 1929 was transferred to Dallas, Tex., in the same position. Returning to the office of the general plant superintendent in St. Louis in 1933, he served in the office of the general plant superintendent until in 1936 he was made area plant supervisor. After serving as outside plant engineer in New York from 1937 to 1940, he became division plant superintendent in St. Louis.

James Mowton Saunders Waring (A '06) retired member of the engineering firm of Chase and Waring, New York, N. Y., died October 23, 1946. He was born January 8, 1874, in Baltimore, Md., and was graduated from Johns Hopkins University in 1896. After experience with the United Railways and Electric Company, Baltimore, from 1896 to 1901 and later with the Electric Storage Battery Company, Chicago, Ill., he joined the firm of L. L. Summers and Company, consulting engineers of Chicago, Ill., about 1911. During the first world war he served as a colonel in the Ordnance Department at Nitro, W. Va. He became a partner in the firm of Chase and Waring in 1927. In 1939 Mr. Waring was appointed research director of the political philosophy and social science department at Fordham University, New York. During World War II he was adviser to the War Production Board and the New York State and Maryland Planning Commissions. He was a former president of the Virginia Military Institute Alumni Association and belonged to the Army and Navy Club of Washington, D. C.

Hector Tabossi (A '12, M '19) director and manager, electrical department, General Electric Company of South America, Buenos Aires, Argentina, died June 15, 1946. Born October 21, 1889, in Concepcion del Uruguay, Argentina, Mr. Tabossi was graduated from Ohio State University in 1910 with the degree of mechanical engineer in electrical engineering. In 1912 he received the degree of master of electrical engineering from Harvard University. After a year with the General Electric Company, Lynn, Mass., he became assistant electrical engineer with J. G. White and Company, Ltd., London, England. In 1916 he joined the British Thomson-Houston Company, Ltd., Rugby, England, as commercial engineer, and remained with it until in 1930 he was named commercial engineer for the General Electric Company of South America. He was appointed director and chief engineer in 1935, and director and manager of the electrical department in 1938.

Ferris LeRoy Francisco (A '04, M '16) senior partner, Francisco and Jacobus, Engineers and Architects, New York, N. Y., died October 11, 1946. Mr. Francisco was born October 23, 1879, in Cleveland, Ohio. His earliest engineering experience was gained at electric light, railway, and power house construction work from 1896 to 1902. In 1902 he became electrical engineer for the American Tobacco Company, Louisville, Ky., and in 1903 he was named supervising electrical engineer. He was appointed chief engineer for the company in 1905. In 1912 he organized his own consulting firm with F. R. Jacobus as partner. Mr. Francisco was a member of the American Society of Civil Engineers, The American Society of Mechanical Engineers, and of the Union League Club of New York.

James Harrison (A '03, F '25) retired engineer of the Southwestern Bell Telephone Company, St. Louis, Mo., died June 16, 1946. Mr. Harrison was born September 25, 1874, in St. Louis, and after attending Washington University for a brief period was graduated in 1896 from Harvard University with the degree of bachelor of science in electrical engineering. He joined the Kinloch Telephone Company, St. Louis, in 1897 as timekeeper and surveyor on conduit work and in 1898 was named cable inspector. The following year he was made wire chief and assistant chief engineer, and in 1906 he was appointed chief engineer. He assumed the additional duties of assistant general manager in 1917. When the company was merged with the Southwestern Bell Telephone Company in 1923, he was titled engineer in the general engineering department. He retired from the Bell company in 1933. During the war he was active in the Victory Loan drives.

Clarence Ray Herrington (A '45) resident engineer, Southwestern Public Service Company, Amarillo, Tex., died September 5, 1946. He was born May 10, 1900, in Kansas City, Mo. Mr. Herrington had been employed by the Public Service company since 1917. With previous experience in meter and relay testing, switchboard wiring, and substation and power plant construction and maintenance, he was made foreman of power plant and substation construction and maintenance in 1939. He became supervisor of substation and power plant maintenance in 1942 and resident engineer at Carlsbad, N. Mex., in 1944. He was a member of the Southwestern Metermen's Association.

Milton P. Kerr (M '41) assistant general manager, Wheeling (W. Va.) Electric Company, died October 7, 1946. He was born January 4, 1889, in Allegheny, Pa. After brief employment as timekeeper and assistant superintendent of the West Penn Electric Company, Pittsburgh, Pa., he was employed as material procurement man and record clerk by the Wheeling Electric Company in 1914. He served in the United States Army in 1918 and returned in 1919 to the Wheeling company as assistant to the general superintendent. He was successively construction superintendent and transmission superintendent for the Ohio Power Company, Canton, from 1923 to 1935. He returned to the Wheeling company in 1936 as general superintendent and recently was appointed assistant general manager.

Ranjit Singh Jain (M '39) professor of electrical engineering, Benares Hindu University, Benares, India, died in April 1946. Professor Jain was born October 3, 1892, in Delhi, India, and was graduated from the University of Illinois with the degree of bachelor of science in electrical engineering in 1915. He was assistant engineer with the Reliance Electric and Engineering Company, Cleveland, Ohio, from 1915 to 1919 and in the latter year was appointed manager of engineering for India. He returned to India in 1921 and was appointed professor at Benares Hindu University. He was acting head of the university in 1939. Professor Jain was an Associate of the AIEE from 1916 to 1937 and was a member of the Institution of Electrical Engineers.

Roy Clarence Arter (A '39) chief engineer, North Electric Manufacturing Company, Galion, Ohio, died September 22, 1946. He was born November 13, 1889, in Galion, Ohio, and attended Ohio Northern University. Mr. Arter entered the employ of the North Electric Company in 1911 as wireman, tester, and draftsman. He was named resident factory engineer in 1914 and engineer in charge of design, production, and sales in 1918. In 1922 he was appointed chief engineer.

MEMBERSHIP • •

Recommended for Transfer

The board of examiners, at its meeting of October 17, 1946, recommended the following members for transfer to the grade of membership indicated. Any objection to these transfers should be filed at once with the secretary of the Institute.

To Grade of Fellow

Bushman, A. K., manager, application & serv. engg. div., apparatus dept., General Elec. Co., Schenectady, N. Y.
Concordia, C., analytical engineer, central station engr. divs., General Elec. Co., Schenectady, N. Y.
Edwards, M. A., assistant engineer, general engg. & const. lab., General Elec. Co., Schenectady, N. Y.
Foust, C. M., division engineer, high voltage & nucleonics div., General Elec. Co., Schenectady, N. Y.
Johnson, E. E., manager of engineering, apparatus dept., General Elec. Co., Schenectady, N. Y.
Kilbourne, C. E., designing engineer, motor & generator div., General Elec. Co., Schenectady, N. Y.
Nance, H. H., engineer, long lines dept., Amer. Tel. & Tel. Co., New York, N. Y.
Suits, C. G., vice-pres. & director, General Elec. Research Lab., Schenectady, N. Y.
White, W. C., electronics engineer, research lab., General Elec. Co., Schenectady, N. Y.
9 to grade of Fellow

To Grade of Member

Agner, O. B., sales engineer, Westinghouse Elec. Corp., Philadelphia, Pa.
AtLee, R. Y., engineer, Amer. District Teleg. Co., New York, N. Y.
Baker, H. W., engineer, service dept., Independent Elec. Machy. Co.; owner, designer engineer, Baker Motion Picture Apparatus Co., Kansas City, Mo.
Bostock, G. M., assistant to mgr., switchgear div., General Elec. Co., Philadelphia, Pa.
Bueche, H. S., prof. electrical engineering, Villanova College, Villanova, Pa.
Buss, R. R., assistant prof. elec. engg., Northwestern University, Evanston, Ill.
Campbell, R. E., engineer, Commonwealth Edison Co., Chicago, Ill.
Chernoff, J. L., electrical engineer, U. S. Bureau of Reclamation, Denver, Colo.
Corden, H. F., general office engineer, dept. of power, Tennessee Valley Authority, Chattanooga, Tenn.
Gross, G. J., assistant engineer, Locke Insulator Corp., Baltimore, Md.
Gund, R. A., product engineer, General Elec. Co., Schenectady, N. Y.
Heckendorn, H., electrical engineer, Western Elec. Co., Chicago, Ill.
Howard, L. E., district cable specialist, General Elec. Co., Philadelphia, Pa.
Hunter, A., electrical designer, Electro-Hydraulics (Messier), Ltd., Warrington, Lanc., England.
Kelley, F. B., supt. electrical construction, Kansas City Pr. & Lt. Co., Kansas City, Mo.
Kurz, C. G., senior electrical engineer, U. S. Maritime Comm., Washington, D. C.
Lebenbaum, P., Jr., electrical engineer, General Elec. Co., Lynn, Mass.
Loustaunau, J. J., chief electrical engineer, E. B. Badger & Sons, Co., Boston, Mass.
Lukens, A. F., industrial engineer, General Elec. Co., Lynn, Mass.
Lutz, S. G., chairman, dept. of elec. engg., New York University, New York, N. Y.
McCartney, C. B., consulting engineer, Box 1050, St. Petersburg, Fla.
Miller, N. C., instructor, University of Minnesota, dept. of elec. engg., Minneapolis, Minn.
Miller, R. E., research & development engineer, Curtis Development & Mfg. Co., Milwaukee, Wis.
Olmsted, F. A., electrical engineer, Erik Floor & Associates, Chicago, Ill.
Pierce, R. M., vice-president in charge of engineering, WGAR, WJR, KMPC, Cleveland, Ohio.
Preisel, E. A., electrical engineer, U. S. Navy Yard, New York, N. Y.
Redding, F. A., supervisor, meter section, frequency change dept., Southern Calif. Edison Co., Ltd., Alhambra, Calif.
Reichel, A. N., chief estimator, Busch Bros., Inc., Englewood, N. J.
Richards, K. W., electrical engineer, Dow Chemical Co., Midland, Mich.
Roos, W. B., supt., laboratory, Mexican Lt. & Pr. Co.; prof. elec. engg., Mexican National University, Mexico, D. F., Mex.
Sanger, J. H., electrical engineer, Walter Walkling & Associates, Denver, Colo.
Springer, C. B., assistant engineer, Locke Insulator Corp., Baltimore, Md.
Stefanetti, H. J., assistant supervisor, Pacific Gas & Elec. Co., Marysville, Calif.
Stobbe, J. A., consultant, 30 Broad St., New York, N. Y.
Torian, J. T., member of technical staff, Bell Tel. Laboratories, Inc., New York, N. Y.
VanWambeck, S. H., associate professor, dept. of elec. engg., Washington University, St. Louis, Mo.
Vowels, R. E., State Electricity Comm., Parbury House, Brisbane, Queensland, Australia.
Williams, C. L., chief engineer, Compania Colombiana de Electricidad, Barranquilla, Colombia, S. A.
Williams, E. M., associate prof. elec. engg., Carnegie Institute of Technology, Pittsburgh, Pa.
Yates, J. E., safety supervisor, Pacific Pr. & Lt. Co., Portland, Oreg.
Yavitz, M., engineer, S. C. Sachs Co., Inc., St. Louis, Mo.
41 to grade of Member

Applications for Election

Applications have been received at headquarters from the following candidates for election to membership in the Institute. Any member objecting to the election of any of these candidates should so inform the secretary before December 21, 1946, or February 21, 1947, if the applicant resides outside of the United States or Canada.

To Grade of Member

Abbett, G. W., Abbett Elec. Co., San Francisco, Calif.
Abray, C. H., Canada Coach Lines, Ltd., Hamilton, Ont., Canada.
Adams, F. H., J. S. Clark Equip. & Mfg. Co., El Paso, Tex.
Allen, M., Messrs. Ashanti Goldfields Corp., Ltd., London, England.
Annis, R. B., R. B. Annis Co., Indianapolis, Ind.
Auten, L. D., R.D. #3, Fredericktown, Ohio.
Ayers, O. T., Jr., Florida Pr. & Lt. Co., Miami, Fla.
Bisbee, G. A. W., E. I. duPont de Nemours, Wilmington, Del.
Bostwick, W. G., Precision Welder & Machine Co., Cincinnati, Ohio.
Brewer, N. R., Sinclair Refining Co., East Chicago, Ill.
Byrd, W., Jr., General Cable Corp., Boston, Mass.
Crider, J. M., U. S. Naval Ship Yard, Charleston, S. C.
Crossan, T. E., Virginia Elec. & Pr. Co., Richmond, Va.
Cutler, C. W., General Elec. Co., Chicago, Ill.
Derry, W. H., Aluminum Co. of America, New Kensington, Pa.
Dunn, R. E., Sanderson & Porter, Saltville, Va.
Fitch, L. C., Public Service Elec. Co., Irvington, N. J.
Forbes, H. E., Alabama Pr. Co., Tuscaloosa, Ala.
Garrison, E. W., Jr., Usco Pr. Equipment Corp., Birmingham, Ala.
Geohegan, W. A., Cornell Univ. Medical College, New York, N. Y.
Green, R. E., West Va. Pulp & Paper Co., Charleston, S. C.
Greene, B. F., 225 W. 34th St., New York, N. Y.
Herr, D. L., Control Instrument Co., Inc., Brooklyn, N. Y.
Humphrey, G. W., B. F. Goodrich Chemical Co., Louisville, Ky.
Iberg, E., The Austin Co., Cleveland, Ohio.
Kist, C., Dept. of Water & Pr., Los Angeles, Calif.
Lovell, W. E., Univ. of Saskatchewan, Saskatoon, Sask., Canada.
McIntire, M. M., Bureau of Reclamation, Phoenix, Ariz.
Mercer, W. R., Raytheon Mfg. Co., Waltham, Mass.
Mitchell, J. G. (Re-election), 7720 Sheridan Ave., Chicago, Ill.
Rawlinson, R., Campbell & Isherwood, Ltd., Liverpool, England.
Richart, R. R., McGraw-Hill Publishing Co., New York, N. Y.
St. John, L. E., General Aniline & Film Corp., Binghamton, N. Y.
Schoenhaar, L. H., Consolidated Gas & Elec. Co., Baltimore, Md.
Simpson, A. E., 850 West Hastings St., Vancouver, B. C., Canada.
Spurgeon, S. J., Continental Gin Co., Birmingham, Ala.
Srinivasan, A., Illinois Institute of Technology, Chicago, Ill.
Stocker, O., Kellogg Co., Battle Creek, Mich.
Stoll, A. L., Toledo Edison Co., Toledo, Ohio.
Strickland, H. A., Jr., The Budd Co., Detroit, Mich.
Swanson, D. H., Line Material Co., Milwaukee, Wis.
Taylor, C. E., Delaware Pr. & Lt. Co., Wilmington, Del.
Taylor, P., Central Pr. & Lt. Co., Corpus Christi, Tex.
Thompson, C. F., Sperry Gyroscope Co., Great Neck, N. Y.
Trainor, J. J., Public Service Co. of Ind., Inc., Indianapolis, Ind.
Tyler, R. L., Central Arizona Lt. & Pr. Co., Phoenix, Ariz.
Trier, R. T., Superior Switchboard and Devices Co., Canton, Ohio.
Twining, G. E., Quinton Engineers, Ltd., Los Angeles, Calif.
Ulrich, V. K., Hytron Radio & Electronics Corp., Salem, Mass.
Waldrip, L. H., L. H. Waldrip Co., Cleveland, Ohio.
Warren, C. L., U. S. Army Engineers, Seattle, Wash.
Williams, J. E., Irvington Varnish & Insulator Co. of Canada, Ltd., Hamilton, Ont., Canada.
Williamson, G. P., Crocker-Wheeler Elec. Mfg. Co., Ampere, N. J.
Wright, N. N., Ferranti Elec., Montreal, Que., Can.
Yevick, J. G., Potomac Elec. & Pr. Co., Washington, D. C.
Zavales, C. T., Westinghouse Elec. Corp., Baltimore, Md.
56 to grade of Member

To Grade of Associate

United States and Canada

1. North Eastern

Baschnagel, R. W., Rochester Gas & Elec. Corp., Rochester, N. Y.
Bence, M. W., Int'l General Elec. Co., Inc., Schenectady, N. Y.
Beuerman, R. A., DuPont Co., Buffalo, N. Y.
Bourdon, A. A., Norton Co., Worcester, Mass.
Estes, J. V., Westinghouse Elec. Corp., South Boston, Mass.
Evans, L. W., Naval School (General Line), Newport, R. I.
Gleason, G. F., Rochester Gas & Elec. Corp., Rochester, N. Y.
Hudgings, D. W., III, Stromberg-Carlson Co., Rochester, N. Y.
Johnson, W. G., Westinghouse Elec. Corp., Springfield, Mass.
Joshi, M. G., General Elec. Co., Pittsfield, Mass.
Klarman, K. J., Union College, Schenectady, N. Y.
Mazur, V. N., Westinghouse Elec. Co., Buffalo, N. Y.
McDonald, J. M., Jr., General Elec. Co., Pittsfield, Mass.
Mellichamp, L. R., Jr., General Elec. Co., Bridgeport, Conn.
Miller, A. E. (Re-election), Westinghouse Elec. Corp., New Haven, Conn.
Offensend, L. D., Eastman Kodak Co., Rochester, N. Y.
Rich, M. (Re-election), Chas. T. Main, Inc., Boston, Mass.
Skalnik, J. G., Yale University, New Haven, Conn.
Stallings, H. B., Clark Controller Co., Buffalo, N. Y.
Talbot, C. G., General Elec. Co., Schenectady, N. Y.
Theleen, D. J., General Elec. Co., Schenectady, N. Y.
Wallace, J. T., A. S. Hamilton, Jr., Consulting Engineers, Rochester, N. Y.

2. Middle Eastern

Adams, H. W., Rural Electrification Admn., Washington, D. C.
Barnette, G. L., Hercules Powder Co., Wilmington, Del.
Bogle, E. R., Jr., General Elec. Co., Philadelphia, Pa.
Bovenizer, W. N., Line Material Co., Zanesville, Ohio.
Carlson, M. E., General Elec. Co., Erie, Pa.
Cropley, W. D., The Hoover Co., North Canton, Ohio.
Crouch, W. G., AF, War Dept., Wright Field, Dayton, Ohio.
De Wyer, G. A., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Dobson, D. R., Stackpole Carbon Co., East Pittsburgh, Pa.
Fagge, W. E. (Re-election), P. O. Box 35, Paoli, Pa.
Felton, W. W., The Franklin Institute, Philadelphia, Pa.
Fesler, J. F., Carbide & Carbon Chem. Corp., S. Charleston, W. Va.
Gerrish, D. I., E. I. duPont de Nemours, Wilmington, Del.
Goessel, E., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Goggins, J. W., Design Service Co., Cleveland, Ohio.
Harr, J. L., Koppers Co., Inc., Pittsburgh, Pa.
Heath, B. W., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Jalen, H. J., Elect. Controller & Mfg. Co., Cleveland, Ohio.
Jokl, A., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Kelch, R. V., The Ohio Power Co., Canton, Ohio.
Kuhn, M. S., Jr., American Ship Building Co., Cleveland, Ohio.
Landes, L. G., Natl. Advisory Comm. for Aeronautics, Cleveland, Ohio.
Larr, R. B., North Elec. Mfg. Co., Galion, Ohio.
Lucal, C. L., The Ohio Pr. Co., Canton, Ohio.
Malick, F. S., Westinghouse Elec. Corp., Pittsburgh, Pa.
McKelvey, W. O., Koppers Co., Inc., Pittsburgh, Pa.
McMillen, J. R., Ohio Pr. Co., Canton, Ohio.
McMillon, A. B., Jr., General Elec. Co., Erie, Pa.
Miller, W. D., Sperry Gyroscope Co., Inc., Philadelphia, Pa.
Nelson, P. C., Westinghouse Elec. Corp., East Pittsburgh, Pa.
Niemi, W., Leece-Neville, Cleveland, Ohio.
Olin, J. R., General Elec. Co., Erie, Pa.
Parks, G. H., Jr., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Pascoe, R. J., Westinghouse Elec. Corp., East Pittsburgh, Pa.
Roth, L. M., Carbide & Carbon Chem. Corp., Charleston, W. Va.
Royston, W. W., Air Materiel Command, Wright Field, Dayton, Ohio.
Schurr, C. A., Elec. Controller & Mfg. Co., Cleveland, Ohio.
Seevers, G. E., The Ohio Pr. Co., Canton, Ohio.
Spicer, G. W., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Sultzbach, R. L., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Thacker, H. B., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Thistle, J. R., Leeds and Northrup Co., Phila. Pa.
Tung, I-J., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Westhoefer, D. L., U. S. Naval Ordnance Plant, Canton, Ohio.
Williams, C. M., Ohio Power Co., East Canton, Ohio.
Williams, T. J., The Ohio Power Co., Canton, Ohio.

3. New York City

Armour, B. J., 1201 University Ave., New York, N. Y.
Barton, C. A., U. S. Rubber Co., New York, New York, N. Y.
Berman, R., Communication Measurements Lab., New York, N. Y.
Brown, L. C., Westinghouse Elec. Corp., Bloomfield, N. J.
Fleisher, M., 278 Legion Street, Brooklyn, N. Y.
Giantvalley, J. R., AIEE, New York, N. Y.
Hames, G. J., Cutler-Hammer, Inc., New York, N. Y.
Hauser, R. C., Ebasco Services, Inc., New York, N. Y.
Hayne, C. A., Int'l Standard Elec. Corp., New York, N. Y.
Hill, D. M., Chemical Construction Corp., New York, N. Y.
Lattin, B. C., 3 Bretton Road, Scarsdale, N. Y.
Margolis, H. B., American Gas & Elec. Service Corp., New York, N. Y.
Merer, G. E., Watson Laboratories, Red Bank, N. J.
O'Neill, B. J., General Aviation Equipment Co., Inc., New York, N. Y.
Polak, A. H., Consolidated Edison Co. of N. Y., New York, N. Y.
Rubin, M. (Re-election), 66 Court St., Brooklyn 2, N. Y.
Schiff, N., 858 Faile St., New York, N. Y.
Schwartz, G. J., Arma Corp., Brooklyn, N. Y.
Shapiro, H., Sanderson & Porter, New York, N. Y.
Slezak, S. T. (Re-election), Bell Tel. Lab., Inc., New York, N. Y.
Smelile, L., 6 Burns St., Forest Hills, N. Y.
Trenkamp, A., Jr., 39 Park Ave., Maplewood, N. J.
Watt, G. J., Sperry Gyroscope Co., Garden City, N. Y.

4. Southern

Anest, N., TVA, Chattanooga, Tenn.
Bradley, J. A., Jr., TVA, Knoxville, Tenn.
Ellis, C. J., General Elec. Co., Chattanooga, Tenn.
Hammer, W. A., Jr., TVA, Chattanooga, Tenn.
Hengy, G. C., Geo. C. Hengy & Co., Shreveport, La.
James, W. F., Usco Pr. Equipment Corp., Birmingham, Ala.
Kirkland, J. W., Gulf States Utilities Co., Lake Charles, La.
Lear, W. E., University of Alabama, University, Ala.
Meier, J. A., Florida Elec. Supply, Inc., Tampa, Fla.
Moore, E. V., Sr., Farnsworth Bldg. Co., Memphis, Tenn.
Riall, E. C., Jr., Southwestern Gas & Elec. Co., Shreveport, La.
St. Dizier, H. A., Gulf States Utilities Co., Lake Charles, La.
Taylor, J. D., General Elec. Co., New Orleans, La.
Van Os, J. H., Tulane University, New Orleans, La.
Wright, E. C., Cities Service Refining Corp., Lake Charles, La.
Wright, R. F., Westinghouse Elec. Corp., Chattanooga, Tenn.

5. Great Lakes

Bloom, C. J., Illinois Bell Tel. Co., Chicago, Ill.
Casey, J. E., Leeds & Northrup Co., Chicago, Ill.
Colip, C. D., Colip Bros., Inc., South Bend, Ind.
Cook, L. D. (Re-election), Commonwealth Edison Co., Chicago, Ill.
Curtis, C. L., Iowa State College, Ames, Iowa.
Damidovich, J. F. (Re-election), Western United Gas & Elec. Co., Aurora, Ill.
Fleming, R. T., General Elec. Co., Detroit, Mich.
Goss, W. R., General Elec. Co., Fort Wayne, Ind.
Hari, A., Western United Gas & Elec. Co., Aurora, Ill.
Hickman, G. K., Wisconsin Elec. Pr. Co., Milwaukee, Wis.
Holcomb, W. L., American Tel. & Tel. Co., Chicago, Ill.
Kepper, C. J., Square D Co., Milwaukee, Wis.
Knuth, N. W., Western United Gas & Elec. Co., Aurora, Ill.
Little, M. C., Western United Gas & Elec. Co., Aurora, Ill.
Loye, J. S., Wood Conversion Co., Cloquet, Minn.
McGuire, J. V., Allis-Chalmers Mfg. Co., Milwaukee, Wis.
McKee, J. L., Ill. Bell Tel. Co., Chicago, Ill.
Reed, E. B., Weltronic Co., Detroit, Mich.
Schultz, W. F., Redmond Co., Inc., Owosso, Mich.
Shefveland, A. M., Public Service Co. of No. Ill., Maywood, Ill.
Sigmund, E. F., Allen-Bradley Co., Milwaukee, Wis.
Singh, P., Purdue University, Lafayette, Ind.
Smith, J., Berwind Fuel Co., Superior, Wis.
Srinivasan, S., 1414 East 59 St., Chicago, Ill.
Stout, E. R., 221 East 43rd Ave., Gary, Ind.
Topczewski, E. A., Cutler-Hammer, Inc., Milwaukee, Wis.
Tunell, L. E., Public Service Co. of No. Ill., Joliet, Ill.
Whittaker, E. C., Sangamo Elec. Co., Springfield, Ill.
Wiprud, R. B., Westinghouse Elec. Corp., Duluth, Minn.

6. North Central

Eyer, J. M., Bureau of Reclamation, Denver, Colo.
Larson, N. M., U. S. Bureau of Reclamation, Denver, Colo.
Law, R. K., U. S. Bureau of Reclamation, Denver, Colo.
Lawrence, W. D., Jr., U. S. Bureau of Reclamation, Denver, Colo.
Peterson, H., Peterson Co., Denver, Colo.
Roe, R. A., Continental Oil Co., Casper, Wyo.
Sinclair, K. M., U. S. Bureau of Reclamation, Denver, Colo.

7. South West

Batterson, C. C., Lt., 4413 Rusk Ave., Houston, Tex.
Craig, R. M., Gulf States Utilities Co., Beaumont, Tex.
Duesterhoeft, W. C., Jr., University of Texas, Austin, Tex.
Emfinger, N. W., Texas Elec. Service Co., Ft. Worth, Tex.
Hassinger, C. V., LCRA, Austin, Tex.
Hobusch, W. G., Jr., Moloney Elec. Transformer Co., St. Louis, Mo.
Lewis, J. P. (Re-election), Aluminum Co. of America, Kansas City, Mo.
Loper, J. F., LCRA, Austin, Tex.
Slade, W. C., Jr., Southwestern Public Service Co., Denver City, Tex.
Sloan, C. E., LCRA, Austin, Tex.
Smith, H. W., The University of Texas, Austin, Tex.
Steele, E. S. (Re-election), Dept. of Public Works, City Hall, Ft. Worth, Tex.
Voter, E. G., Southwestern Bell Tel. Co., Dallas, Tex.
Wanja, L. F., Texas Pr. & Lt. Co., Dallas, Tex.
Woods, K. H., Southwestern Public Service Co., Plainview, Tex.

8. Pacific

Breuer, J. R., General Elec. Co., San Francisco, Calif
Conrey, D. W. (Re-election), Southern Calif. Tel. Co., Los Angeles, Calif.
Coombs, G. F., California Elec. Works, San Diego, Calif.
Cornell, L. P., Jr., The Pacific Tel. & Tel. Co., San Francisco, Calif.
Dalzell, L. B., Southern Calif. Tel. Co., Los Angeles, Calif.
Devine, G. V., The Pacific Tel. & Tel. Co., San Francisco, Calif.
Dixon, A. A., Modesto Irrigation District, Modesto, Calif.
Dolly, W. W., Supervisor of Ships Office, USN, Terminal Island, Calif.
Emerson, F. E., Railroad Comm. of the State of Calif., San Francisco, Calif.
Hall, R. V., Pacific Gas & Elec. Co., Fresno, Calif.
Lindquist, W. W., Pacific Gas & Elec. Co., Oakland, Calif.
Mahler, F. C., J. E. Redmond Supply Co., Phoenix, Ariz.
McKay, H. B., The Pacific Tel. & Tel. Co., San Francisco, Calif.
McMurtry, D. P., U. S. AAF, Fairfield Army Air Base, Calif.
Raab, W. J., Calif. Prune & Apricot Growers Assoc., San Jose, Calif.
Riha, A. E., Pacific Elec. Mfg. Corp., San Francisco, Calif.
Roadman, C. O., Central Arizona Lt. & Pr. Co., Phoenix, Ariz.
Seaman, R. G., Bechtel Bros. & McCone, Hunting Park, Calif.
Sherwood, H. L., Pacific Gas & Elec. Co., Oakland, Calif.
Sipe, H. T., Pacific Gas & Elec. Co., Oakland, Calif.
Slater, J. E., Stolper Elec. Co., Burbank, Calif.
Stand, T. R., Station KFRC, San Francisco, Calif.
Tarratt, H. P., County of San Mateo, Court House, Redwood City, Calif.
Thomas, R. E., Weihe, Frich & Kruse, San Francisco, Calif.

9. North West

Black, R. C., General Elec. Co., Seattle, Wash.
Dole, R. S., Clearwater Valley Lt. & Pr. Assoc., Lewiston, Idaho.
Loring, A. E., U. S. Government, Fort Douglas, Utah.

10. Canada

Garratt, C., Ferranti Elec. Ltd., Mount Dennis Toronto, Ont., Canada.
Houlding, J. D., Canadian Westinghouse Co., Hamilton, Ont., Canada.
Painter, G. W., Canadian General Elec. Co., Ltd., Toronto, Ont., Canada.
Sorby, W. O., Canadian Westinghouse Co., Ltd., Montreal, Que., Canada.

Elsewhere

Chinea, C., R., Nicaro Nickel Co., Nicaro, Oriente, Cuba.
Corbin, G. F., Ketchikan Public Utilities, Ketchikan, Alaska.
Ingram, G. W., Shell Petroleum Co., Ltd., London, England.
Pettit, R. D., The British Thomson-Houston Co., Ltd., Rugby, England.
Toepfer, H. J., Shanghai Pr. Co., Shanghai, China.
Wanger, W., Brown, Boveri & Co., Ltd., Baden, Switzerland.

Total to grade of Associate
United States and Canada, 189
Elsewhere, 6

OF CURRENT INTEREST

Second National Electronics Conference Held in Chicago, October 3-5

With an advance registration of about 1,200 and a final registration of approximately 2,100, the second National Electronics Conference was held October 3–5, 1946, at the Edgewater Beach Hotel in Chicago. During the $2^1/_2$-day conference 19 technical sessions—running as many as 5 in parallel—accommodated 57 lists of technical subjects which were presented in various forms ranging from formal technical papers to extemporaneous reports and panel discussions.

A popular special feature of the conference program was the series of equipment exhibits and demonstrations which filled two large halls in the portion of the building set aside for conference activities. Other special features of the program included two huge general luncheons, one addressed by President Frederick L. Hovde of Purdue University on the subject, "Science, Politics, and the National Welfare"; the other of which was addressed by Vice-President C. G. Suits (M '41) of the General Electric Company on the subject, "Physics of Today Becomes the Engineering of Tomorrow." President Hovde's address will be found on pages 554–6 and that of Doctor Suits will appear in an early issue of *Electrical Engineering*. A joint meeting of the Chicago Sections of the AIEE and the IRE was held Friday evening in collaboration with the National Electronics Conference, at which Doctor J. O. Perrine, assistant vice-president of the Bell Telephone Laboratories, Inc., presented one of his notable animated demonstration lectures "Radar and Microwaves."

1947 CONFERENCE ANNOUNCED

Preliminary plans for the third National Electronics Conference already have been completed and the announcement made that the conference will be held at the Edgewater Beach Hotel in Chicago, November 3–5, 1947. Tentative plans also were announced for the fourth conference, to be held at the same place and at approximately the same date in 1948. It is expected that these future conferences will be of the same nature as the two already held, except that additional attention and probably additional space will be devoted to equipment exhibits and demonstrations. It is the announced plan and objective of the conference sponsors that the program continue to offer an industry-wide opportunity for the interim reports and discussion of any current electronics subject matter regardless of the specialized subdivision of the electrical field, from which the information may become available.

On the basis of present scheduling, the AIEE National Midwest Meeting will be held at the Congress Hotel in Chicago the week of November 3–7, 1947. It is expected that the programs of these two important and somewhat interrelated meetings will be appropriately co-ordinated and correlated, so that AIEE members and others interested in attending one of them will have the opportunity also to take advantage of the extended range of subject matter offered by the other.

ORIGIN OF NEC

The organization of the first National Electronics Conference held in Chicago, October 5–7, 1944, was the outgrowth of the initiative and imagination of a group of ten men who met March 23, 1944, to consider some of the needs and implications growing out of the wartime mushroom growth in the field of applied electronics and related research. The group included P. G. Andres, T. J. Higgins (M '46) J. E. Hobson (M '41), C. S. Roys (M '45), and E. H. Schulz (M '43), all of the Illinois Institute of Technology; R. E. Beam (A '42), A. B. Bronwell (M '44), J. F. Calvert (F '45), all of Northwestern University; Beverly Dudley (M '43) of the McGraw-Hill Publishing Company; and W. O. Swinyard of the Hazeltine Electronics Corporation. Out of the decisions of this group came the first National Electronics Conference, financially underwritten jointly by Northwestern University and Illinois Institute of Technology.

Subsequently, the conference was incorporated under the laws of the state of Illinois as a nonprofit organization: "Serving as a national forum for the presentation of authoritative technical papers on electronic research, development, and application." The officers of National Electronics Conference, Inc., for 1946 are:

President W. O. Swinyard, Haseltine Electronics Corporation
Executive Vice-President A. B. Bronwell, (M '44) Northwestern University
Vice-President R. E. Beam, (A '42) Northwestern University
Vice-President C. A. Emery, Westinghouse Electric Corporation
Vice-President L. T. Rader (M '43) Illinois Institute of Technology
Treasurer W. M. Ballenger (M '38) General Electric Company
Secretary E. H. Schulz, (M '43) Illinois Institute of Technology

In addition to the foregoing group of officers, the NEC board of directors includes the following:

J. E. Hobson (M '41) Armour Research Foundation
Alfred Crossley Consulting Engineer
R. J. Donaldson (M '44) Commonwealth Edison Company
G. H. Fett (M '38) University of Illinois
R. H. Herrick, Automatic Electric Company
L. S. McPhee, Whiting Corporation
Cullen Moore, Galvin Manufacturing Company
E. O. Neubauer (M '44) Illinois Bell Telephone Company
C. S. Roys (M '45) Illinois Institute of Technology
O. D. Westerberg, Illinois Institute of Technology

The policy and objectives of the National Electronics Conference are reflected further in the following excerpts from the "Foreword" of the "Proceedings" of the 1944 Conference.

"To a large extent, the formation of the National Electronics Conference has been the outcome of the recent rapid growth in the field of applied electronics. Supplementing the important role of electronics in communication, the increased importance of electronic methods and devices in physical measurements, medical application, and the rapid growth of industrial application, have focused public attention on the science and technology of electronics.

"Applications of electronics for military communication and various control operations have been an important factor in promoting the popular appeal of this branch of science which is not so young as commonly is supposed. The origin of electronics—the branch of science and technology which relates to the conduction of electricity in gases or in a vacuum—can be traced back to the observation of the unilateral conductivity of an incandescent lamp with enclosed plate, by Thomas A. Edison in 1882 and of the discovery of the electron in 1895 by J. J. Thomson. Important communication applications of electronic-tube amplifiers had been made by 1915, and by 1923 the radiobroadcasting industry was about to be launched. By 1930, feeble attempts had been made to apply electronic devices in industrial uses but another decade was required to overcome objections to "fragile glass tubes" and to establish a sound footing for industrial electronics.

"While the fundamental principles of electronics and electronic devices are well known and adequately recorded (particularly in the communication field), the diverse applications of this branch of technology too often have been developed in specialized divisions of science. This practice has tended to compartmentalize certain advances, and has made more difficult the dissemination of new discoveries and applications outside of that field in which it originated. It had been evident for some time that a new mechanism, correlating the advances in all branches of electronics, would be desirable."

This is the background of the National Electronics Conference, a "National forum on electronics development and their application," now jointly sponsored by the Illinois Institute of Technology, Northwestern University, The University of Illinois, the Chicago Section of the AIEE, and the Chicago Section of the Institute of Radio Engineers, with the co-operation of the Chicago Technical Societies Council.

OPENING SESSION

At the opening session, President H. T. Heald of Illinois Institute of Technology delivered the address of welcome which will be published in a future issue of *Electrical Engineering*, as will the speech of Doctor E. U. Condon (M '44) director of the National Bureau of Standards, who spoke on "Electronics and the Future."

Power-Line Telephone Service Soon Available in Six States

Installation has been started on equipment providing rural telephone service over electric power lines in six states, the American Telephone and Telegraph Company, New York, N. Y., recently announced.

Though telephone subscribers in Alabama and Arkansas have been served temperarily by the new method as an experiment, the new service installations will mark the first time that power-line carrier has been utilized to bring telephone service to rural areas beyond the reach of existing telephone lines.

Five Bell telephone companies, two independent telephone companies, four rural power systems financed by the Rural Electrification Administration, and three power companies are parties to the installations. Telephone central offices through which the seven groups of rural subscribers will be served are situated at Manakin, Va.; Nashville, N. C.; Aiken, S. C.; Italy and Lamesa, Tex.; Oak Creek, Colo.; and Cle Elum, Wash.

Equipment for the installations has been manufactured by the Western Electric Company which also has commenced production to meet the expected extension of this type of service.

The telephone company also revealed that an experimental project is scheduled for this fall at Norton Mills, Vt.

Though the projects now in process of installation will be equipped to provide only one telephone channel for each power-line route, power-line carrier apparatus now under development is expected to furnish six speech channels.

Patent Examiners Needed by the United States Patent Office

The United States Patent Office has announced that expansion of its examining corps has created a need for additional patent examiners in grades P-1 and P-2 with starting salaries of $2,644.80 and $3,397.20 respectively.

To qualify for an appointment to a P-1 patent examining position, applicants must have completed a full curriculum of study leading to a bachelor's degree in a college or university of recognized standing including or supplemented by major study in engineering, technology, chemistry or physics; for four years' experience in one of the foregoing fields of such a nature as to demonstrate that the applicants possess an intimate working knowledge of the field involved to the same extent and degree that such a knowledge would have been acquired through college work. To qualify for appointment to a P-2 patent examining position, in addition to the requirements for a P-1 appointment, applicants must show at least one year of responsible experience in the field of chemistry, physics, technology, engineering, or other pertinent work.

After six months of satisfactory service in the P-1 grade, patent examiners are eligible for promotion to grade P-2. Promotions to the higher grades are made from within the examining corps when the employee meets the qualifications for the higher grades.

All interested persons may obtain a Civil Service Commission application form 57, at any first- or second-class Post Office, or Civil Service Commission regional office. Applications should be mailed to the United States Patent Office, Personnel Division, Washington 25, D. C.

Magnetron Adapted to Cooking Purposes

The newly developed "Radarange" which uses ultrahigh frequency electromagnetic power to defrost, heat, or cook a wide variety of foods in a matter of seconds recently was demonstrated by the Raytheon Manufacturing Company, Waltham, Mass.

For use by air lines a special model has been developed which weighs approximately 100 pounds and is designed to defrost and cook a complete meal (which previously has been partially precooked and frozen) in less than a minute. The machine operates on about $4^1/_2$ kw from the standard 28-volt d-c power system. The 400-cycle output of motor-generator sets is amplified and rectified to produce direct current at 4,000 volts which is applied to the plate of a continuous-wave magnetron oscillator. The 3,000-megacycle 1,000-watt output of the magnetron is fed to a 4- by 8-inch wave guide by a coaxial cable. The wave guide directs the power to the food which has been placed in a metal tray. The ray acts as a reflector, thereby forcing any stray energy back into the food. Another model for use in drugstores and sandwich shops is ready for production, and a radarange for home use is being developed.

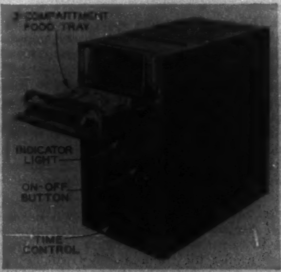


Figure 1. Air line model of the Radarange

This unique cooking device produces hamburgers in 35 seconds. Gingerbread actually can be seen to rise from batter to fluffy muffins in 29 seconds. These muffins have no crusts because the heat is generated in each molecule of the dough and does not have to be conducted from the surface. Hot dogs can be cooked in paper wrappings without burning the paper. This is possible because the water content of the paper is much lower than that of the frankfurter, and water content determines the reaction of each substance to the ultrahigh-frequency energy. As yet no method has been found to boil potatoes, and an egg placed in the range will crack in less than a second. However, it is possible to produce scrambled eggs and baked potatoes.

Portable Instrument Measures Group Opinion

A new instrument designed to measure and indicate the composite opinion of a group of as many as 120 individuals has been announced by the General Electric Company. Called an opinion meter, the instrument enables each person in a group to express automatically, his opinion on any subject. All the individual opinions are summated and registered on a large dial as a single figure in about ten seconds. The instrument has possible use in schools, conference rooms, lecture groups, adult educational organizations, radio stations and political groups.

The opinion meter consists of a large indicating unit and up to 120 hand-held stations. On the individual station is a dial calibrated from 0 to 100, and an adjustable pointer.

The indicating unit is the size of a small suitcase. The cover opens upwards to form a large dial and pointer. Plugs for the string of individual units and for 60-cycle a-c power supply are located on the back of the unit. Inside are electrode and electronic components which perform the measuring and indicating functions.

To register his degree of opinion on a question under discussion, each member of a group turns the pointer on his individual unit to a number on the dial. If strongly in favor of a question, this would be close to 100; if indifferent, 50; and if disapproving, close to 0. The percentage of those in the group who turn their stations to "off," not desiring to express an opinion, also can be determined.

Although the main purpose of the opinion meter is to determine the composite opinion of a group, it also can be used to obtain a ballot. Fifty-fifty opinions are not taken into account in the ballot results, nor are those stations turned to "off."

Exhibitors' Plans Develop for Electrical Exposition

News of exhibits now being developed for the Electrical Engineering Exposition, to be held in New York at 71st Regiment Armory next January 27–31, indicates that the display will open a new channel for information and distribution in a field not hitherto serviced directly by any exposition. The exposition provides an additional feature of interest for those attending the AIEE winter meeting.

For example the introduction of the electronic signal circuit, amplified and relayed to control high voltages, will be featured, and also many new developments in instrumentation specifically applied to the generation and distribution of electric power.

There will be a variety of insulating materials on display, including several classes of plastics, with extensive arrays of products and parts for which they are appropriate, and well-presented information respecting their dielectric properties.

Especially appropriate to this exposition will be the large group of exhibits pertaining to field work, from conductors and pole line equipment of various kinds to transformers, switchgear, and many items of detail—an electronic circuit breaker is one; another is a facsimile demonstration of a lightning arrester at work.

The important needs of maintaining service have not been overlooked, and a number of exhibits will be devoted to maintenance, repair, replacement, and emergency duties, from a newly developed field service truck to a long list of items for in-plant use in time of need.

The armory will be closed to the public, so that a strictly professional and business atmosphere will be maintained. Admission will be by invitation and registration of only qualified visitors.

Medal Presented to Doctor Willis R. Whitney

The fall meeting of the Industrial Research Institute, held October 16–18, at the Westchester Country Club, Rye, N. Y., which brought together more than 100 representatives from the research departments of various industrial organizations, was highlighted by the presentation of the institute's medal to Doctor Willis R. Whitney (A'01) organizer and first director of the General Electric Research Laboratory.

The medal, conferred for the first time since it was established in 1945, was awarded to Doctor Whitney for outstanding contributions to the field of industrial research as a distinguished scientist, a pioneer in the application of organized science to industrial technology, and a beloved and inspiring leader of men. H. W. Graham, senior past president of the Industrial Research Institute, made the presentation at an informal dinner on October 17.

Charles S. Venable, president of the organization, acted as toastmaster at the dinner and introduced the guests at the speakers' table after which Irving Langmuir, associate director of the General Electric Research Laboratory spoke on "Whitney, the Man and Leader." Doctor Langmuir stressed Doctor Whitney's application of serendipity, Webster's definition of which he modified to "the gift of finding *and applying* valuable or agreeable things not sought for," to research. Doctor Whitney, he recalled, always asked those working in the laboratory if they were having fun and never assigned anyone to a project, but let his aides choose according to their own imagination.

In his short address following acceptance of the medal, Doctor Whitney credited Francis Bacon with having started organized research, but on a strictly nationalistic basis with the benefits limited to England alone. Doctor Whitney was emphatic in his opinion that international co-operation among scientists is essential to world peace.

Standards Announced for Fractional-Horsepower Motors

A new program of standardization for fractional-horsepower motors was announced in September by the members of the motor and generator division of the National Electric Manufacturers Association (*EE, Nov '46, p 541*). At that time a basic standard for determining motor rating in co-ordinated terms of horsepower, speed, breakdown torque, and service factor was presented to eliminate certain application difficulties involving rating and performance characteristics.

The standard presented in October provides for uniformity of mounting dimensions and three frame sizes for motors ranging from 1/20 through 3/4 horsepower at 1,800 rpm and including one horsepower at 3,600 rpm. Previously the variety of mounting dimensions and motor sizes necessitated extensive use of many universal mounting bases and adapter plates which resulted in high costs and loss of production time in application of these motors to the products of different manufacturers. The same difficulties have been encountered by service men in attempting to replace a motor of one make with one of another make.

The new dimensions apply to both rigid and resilient or cushion base motors. Those dimensions for which standards have been established include the distance between the axial centerline of the shaft and the bottom of the feet (*D* dimension), distance between the vertical center line and the mounting slot centerlines—end view (*E* dimension), half distance between mounting slot centerlines—side view (*F* dimension), distance from shaft shoulder to nearest mounting hole centerline (*BA* dimension), and diameter of mounting slot or hole (*H* dimension).

In the proposed system for numbering the frames the frame number will be 16 times the shaft height (*D* dimension), and the other dimensions will be constant for any frame size. The three frames will be 42, 56, and 66 with suffix letters used to indicate the same features provided for in the established integral system.

CIGRE Papers Wanted. The Engineering Societies Library has inquired whether any member of AIEE has a set of the papers presented at the 1946 meeting of the International Conference on Large Electric High Voltage Systems, which he would be willing to offer the library. The library has placed an order for a bound volume of the papers, but this will not be published before the end of the year. Correspondence about the papers may be addressed to Ralph H. Phelps, director, Engineering Societies Library, 33 West 39th Street, New York 18, N. Y.

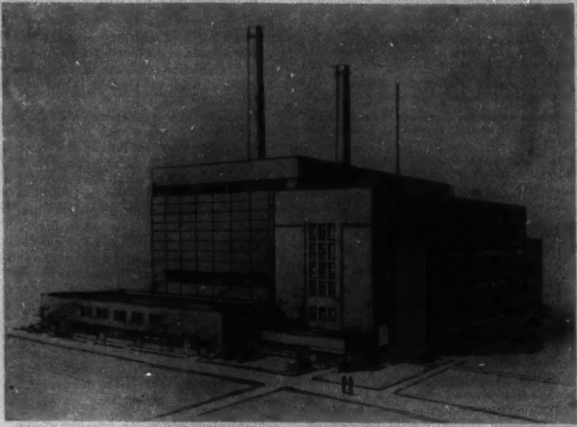
Television at Convention. Television solved the problem of accommodating the overflow audience at the sessions of the 24th annual convention of the National Broadcasters Association held at the Palmer House in Chicago, Ill., recently. Three RCA Image Orthicon television viewing equipments were used to pick up and transmit events in the meeting rooms to 20 RCA Victor television receivers installed in the hotel's exhibition hall for the overflow audience. According to Henry Rhea, manager of RCA television equipment sales, use of the supersensitive Image Orthicon camera permitted pickup with ordinary room lighting, whereas the discomforts, cost, and inconvenience of the special brilliant lighting otherwise required would make such a service impractical.

Commercial Air Line Installs Radar. First radar installations in a $300,000 experimental program recently were made by American Airlines, Inc. The flagship *St. Joseph*, a four-engine DC-4 cargo airplane, became the first commercial airplane to be equipped with radar to facilitate instrument landings in bad weather. A $15,000 radar beacon station is being installed on the air line's hangar in St. Joseph, Mo., and a twin-engine DC-3 laboratory ship has been made available for pilot instruction.

Experts Aid Navy Science. Appointment of ten experts to the newly established Civilian Research Advisory Committee has been announced by Secretary of the Navy James Forrestal. Named to the committee which was authorized by the last session of Congress to assist in spurring Naval scientific research are: Richard J. Dearborn (A'16) president, Texaco Development Corporation, New York, N. Y.; Doctor Karl T. Compton (F'31) president of the Massachusetts Institute of Technology, Cambridge; Lewis L. Strauss, United States Naval Reserve, of Kuhn, Loeb and Company, New York, N. Y.; Luis de Florez, USNR, vice-president in charge of engineering, Doubleday and Company, New York, N. Y.; Doctor Warren Weaver, director of the division of natural sciences, Rockefeller Foundation, New York, N. Y.; Doctor Philip M. Morse, professor of physics, Massachusetts Institute of Technology, who will become director of the Northeastern Regional Laboratory for Atomic Energy Research, Camp Upton, N. Y.; Doctor L. A. DuBridge, president of the California Institute of Technology, Pasadena; Doctor Arthur H. Compton, chancellor of Washington University, St. Louis; Doctor William Sharp McCann, director of the Institute of Medicine, Rochester (N. Y.) University; Doctor Detlev W. Bronk, head of the National Research Council.

Icaroscope Described to Optical Society. The secret of another war-born defensive weapon, the Icaroscope, was revealed by Professor Brian O'Brien (M '28) director of the Institute of Optics of the University of Rochester, N. Y., at the 34th annual meeting of the Optical Society of America held recently in New York. Shaped like a stubby telescope, the Icaroscope enables pilots to find enemy airplanes diving toward them directly in line with the sun. The image of the airplane is formed on a transparent screen made from a phosphor chosen for short afterglow with saturation, while the screen is hidden from the eye. A double rotating screen then closes off the outside light and the pilot sees the image 1/100 second later. By that time the image on the screen is only 20 to 50 times the brilliance of the surrounding sky, as compared with the sun's real brightness of 10,000 to 100,000 that of the surrounding sky, and the airplane is silhouetted on the screen against either the sun or the surrounding sky. The disks are rotated by an electric motor at about 100 cycles per second, so that the images appear with the rapidity of a motion picture. The Icaroscope was developed under contract with the Office of Research and Development as an aid in defense against aircraft attacking from the general direction of the sun. It also was used in observing and photographing the Bikini atom bomb tests.

Generating Station Designed for Rochester



This new $15,000,000 steam electric generating station will be situated on the shores of Lake Ontario at Rochester, N. Y., the Rochester Gas and Electric Corporation has announced. The station, which will have an ultimate capacity of 160,000 kw will be constructed in four units of 40,000 kw each, the first of which is expected to be completed in 1948 at a cost of $8,000,000. Each unit will include one boiler and one turbogenerator with all the necessary auxiliary equipment. The turbine will be operated by steam at 1,250 pounds pressure at the turbine throttle. Station dimensions will be 280 by 250 feet by 100 feet high

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OTHER SOCIETIES •

Building Conference Organizes Building Officials Foundation

Establishment of the Building Officials Foundation, a co-ordinating and research body, which will work for uniform and modern building codes throughout the United States, was announced at the 31st annual meeting of the Building Officials Conference of America, Inc., held recently in Memphis, Tenn.

The principal functions of the Foundation were given as:

1. To encourage the adoption of the "Basic Building Code" promulgated by the Building Officials Conference of America, Inc., by all communities, to the end that a country-wide uniform code be established and to encourage communities to keep their local codes up to date.

2. To assist communities in the administration of their building laws and regulations.

3. To reconcile local building codes with advances in building to the end that structural design, equipment, construction materials, and methods may be progressively improved in the public interest.

4. To provide uniform testing procedure in consultation with industry in determing the adequacy of testing equipment, and for the evaluation of the integrity of equipment, materials, and methods of construction.

5. To make known the results of tests promptly and in a manner compatible with the needs for such data in administering building rules and regulations.

6. To keep building and other municipal officials concerned with public health and safety, manufacturers, producers, architects, engineers, general and special trades contractors and other builders, real estate dealers and administrators, property managers and appraisers, mortage and equity financiers, government officials concerned with construction, building labor leaders, materials dealers, suppliers and distributors and others in or related to the building industry informed of advances in building requirements and code administration and to supply them with news, informative articles, and other data pertaining to better building and construction practices.

The services of the Building Officials Foundation are designed to function primarily for building officials as administrative agents of building laws and regulations, and for the betterment of the building industry as a whole. The various service instruments are as follows:

1. The Basic Building Code, with its Construction Code and Small Municipality Building Code.

2. The structural bureau to analyze plans and projects of unusual magnitude or design submitted by understaffed municipalities.

3. The testing procedure for the introduction of new devices, materials, and methods meeting the functional requirements of building as provided in the Basic Building Code.

4. The *Official Bulletin* of the Building Officials Foundation containing test reports and recommendations for the use of approved materials, equipment, and construction methods.

5. A monthly illustrated magazine for general distribution to building officials and other interested persons in the building industry.

The foundation is a nonprofit organization and was set up by a voluntary endowment fund contributed by the construction industry. For future expenses the foundation will depend on annual dues of participating members and fees collected from

Future Meetings of Other Societies

American Chemical Society. 111th national meeting, April 14–18, 1947, Atlantic City, N. J. 50th annual meeting, June 16–20, 1947, Atlantic City, N. J.

American Society for Testing Materials. Spring meeting and committee week, February 24–28, 1947, Philadelphia, Pa.

American Society of Mechanical Engineers. Annual meeting, December 2–6, 1946, New York, N. Y.

American Society for X-ray and Electron Diffraction. Winter meeting, December 5–7, 1946, Pittsburgh, Pa.

Institute of Radio Engineers. Annual meeting, March 3–7, 1947, New York, N. Y.

National Exposition of Power and Mechanical Engineering. December 2–7, 1946, New York, N. Y.

tests and the promulgation of standards. The grades of membership will be:

1. Ordinary members will include all building officials.
2. Members of the Building Officials Conference of America.
3. Founder members are subscribers to the voluntary endowment fund.
4. Participating members may be manufacturers, producers, financial institutions, insurance companies, general contractors, and others engaged in the building industry.

IES Lighting Handbook Postponed. Issuance of the Illuminating Engineering Society "Lighting Handbook" originally scheduled for 1946 has been postponed to October 1947. In making the announcement, C. A. Atherton, chairman of the handbook committee, explained that reconversion problems have otherwise occupied members working on the handbook, so that production necessarily must be delayed. Postponement of publication is expected to aid manufacturers in supplying the most recent material for the reference data section. When completed, the 500-page handbook will cover every phase of lighting from the pure physics of light to specific lighting recommendations for stores, offices, homes, factories, and even for television studios and juke boxes. The latest information on light sources, as well as the measurement and control of light will be included.

Civil Engineers Nominate President. Edgar M. Hastings of Richmond, Va., chief engineer of the Richmond, Fredericksburg, and Potomac Railroad Company, has been nominated as the 1947 president of the American Society of Civil Engineers. A graduate of Baltimore City College and Baltimore Polytechnic Institute, Mr. Hastings' entire professional career has been in the field of railroad engineering, and he has been associated with the Richmond, Fredericksburg and Potomac Railroad since 1903. He has been chief engineer since 1922. Mr. Hastings is 64 years old.

Relations With Government Considered by Canadian Council

Collective bargaining, the type of technical employment service provided by the government, and school curriculums made up the agenda of the fourth regular meeting of the Canadian Council of Professional Engineers held in Ottawa, Ontario, in September.

The meeting was concerned with the rate at which experienced staff is leaving the service of the Wartime Bureau of Technical Personnel, and with the attempt of the Executive and Professional Division of the National Employment Service to deal with technical persons, with completely unsatisfactory results. It was agreed to direct the attention of the Minister of Labour to these developments, and at the same time to advise the co-operating organizations with respect to the danger of serious deterioration in the quality of national employment service available to their members.

Recent decisions of the National Labour Relations Board on certification of bargaining units of professional people were reviewed, and it was agreed to discuss certain aspects of these decisions on an informal basis with senior administrative officers of the Dominion Department of Labour. The attention of co-operating organizations was directed to the fact that the Minister of Labour had invited representatives of all provinces to meet in Ottawa in an endeavor to draft a national labor code to serve as a basis for provincial labor legislation.

It was announced that information had been assembled regarding all undergraduate and postgraduate scholarships, bursaries, and the like in the field of science at Canadian universities and that the material was being classified for presentation in a final form for the next meeting. A draft of a brief statement outlining the existing collective bargaining situation in Canada as it affects technical persons and professional engineering and scientific organizations was presented to the meeting, and it was decided to transmit copies to all co-operating organizations.

U. S.–Canadian RMA Meeting. United States and Canadian directors of the Radio Manufacturers Association exchanged information on production, labor relations, and school sound equipment at a joint conference held in Quebec, Quebec, in October. A similar joint conference is planned for next spring in a United States city. During the meeting RMA directors appropriated $10,000 to be spent in the promotion of National Radio Week, November 24–30 in conjunction with the National Association of Broadcasters, as the first step toward a closer liason between the two associations. The RMA also voted to support the manufacturer–agent disposal system being used by the War Assets Administration for surplus electronic equipment and to reject a Department of Commerce proposal that the RMA screen captured German technical documents on electronic developments.

AIME Saunders Medal Awarded. The award of the William Lawrence Saunders Gold Medal for 1947 to LeRoy Salsich, president of the Oliver Iron Mining Company, Duluth, Minn., has been announced by the American Institute of Mining and Metallurgical Engineers. The medal was awarded Mr. Salsich for "his conspicuous success in developing men and methods for the mining and transportation of iron ore; for his significant contribution, as operating head of the world's largest iron mining enterprise, to the nation's production of steel so vital to victory in World War II." Mr. Salsich has been associated with the Oliver company, a subsidiary of the United States Steel Corporation, since his graduation from the University of Wisconsin in 1901. He was born in 1879 in Hartland, Wis.

EDUCATION . . .

Cornell Receives Navy Gift. An outright gift of a laboratory unit of Diesel engines and related equipment valued at more than $2,000,000 recently was made by the Navy Department to Cornell University, Ithaca, N. Y. The equipment, accumulated during the war in the course of training almost 3,000 Naval officers, is described as a completely integrated installation adapted to professional instruction and research in the theory, operation, and maintenance of Diesel engines. The main laboratory's 23 engines, the largest weighing 10 tons, range from 1 to 16 cylinders, generating from 15 to 1,700 brake horsepower and represent about 15 different manufacturers. In addition the gift includes the equipment of two auxiliary laboratories, one in which preliminary training in internal combustion was given during the war, and one containing the modern electric equipment used on Naval vessels, as well as air compressors, pumps, refrigeration units, and a communication system. In accepting the gift, Vice-President S. C. Hollister, dean of the college of engineering, said that the equipment not only could be used in existing courses in mechanical and electrical engineering but also would enable the university to organize new elective courses, both graduate and undergraduate.

Georgia School Offers Doctor's Degree. The first engineering and scientific courses in the State of Georgia leading to the degree of doctor of philosophy are being offered by the Georgia School of Technology, Atlanta. The graduate division is scheduling 357 courses which can be pursued by graduate students toward master of science and doctorate degrees. Graduate students applying for the doctor's degree are encouraged to undertake research which will benefit the South from an engineering and industrial viewpoint. The School of Technology also has available financial aids ranging from $600 to $1,800 per academic year to assist students who would be unable to continue advanced studies because of financial need. These include assistantships, instructorships, and fellowships.

Arkansas Gets Engineering Courses. Four-year courses in electrical, mechanical, and civil engineering have been instituted at Arkansas State College, Jonesboro, in response to the current demand for engineering curricula. Applications are being accepted for enrollment in the quarter beginning November 29, 1946, and for the fall semester of 1947. The campus has adequate living accommodations for single men on a barrack basis. New laboratory equipment is being collected and installed for the engineering department. Most of the college buildings have been constructed since 1932, though the college itself was founded in 1909.

INDUSTRY • • • • •

Distribution Department Uses Two-Way Radio Communication

The electric distribution department of the Public Service Electric and Gas Company, Newark, N. J., has completed the construction of a two-way radio communication system by means of which supervisors of the electric operating divisions can communicate directly with line crews, trouble men, and field supervisors wherever they may be working.

Seven fixed stations at different cities in New Jersey serve as centers from which orders are transmitted to field crews in the surrounding territories. Each station has 50-watt frequency-modulated transmitters and antennas more than 100 feet high. Receiving equipment also is installed in the fixed stations.

Mobile radio equipment is being installed in more than 150 cars and trucks. These units are equipped with frequency-modulated receivers and 35-watt transmitters to permit crews to report back. Both receivers and transmitters are energized from the storage battery of the car or truck.

A conventional telephone handset with a "push to talk" button is used at both mobile and fixed stations. As all the equipment is designed for the same frequency, it is possible for any mobile unit to contact another mobile unit.

Utility Council Formed. Formation of the New York State Utility Council was the outcome of a recent two-day meeting of nearly 100 executives of gas and electric companies at Saranac Lake, N. Y. The new council, it was decided, will meet informally from time to time to discuss common problems. Robert E. Ginna, vice-president of the Rochester (N. Y.) Gas and Electric Corporation was designated chairman for the coming year. Among the addresses at the recent meeting was one on "Economic Research as a Basis for Selective Load Building" by Constantine Bary (M '43) rate research engineer, for the Philadelphia (Pa.) Electric Company.

Tube Testing Chamber Converted From War Uses

A tube testing chamber originally designed to test tubes for wartime aircraft equipment now has been equipped for multipurpose testing of all types of tubes, the General Electric Company has disclosed.

Capable of simulating a wide range of temperature and altitude conditions within brief periods of time, the over-all equipment for the chamber occupies a space of approximately seven feet by eight feet by 15 feet. Altitude is attained at the rate of 3,000 feet a minute while 100 degrees below zero can be reached in about two hours and 175 degrees above zero in about 45 minutes. Two dials outside the door automatically control, operate, and record, on a 24-hour chart conditions within the chamber.

Leads into and out of the chamber furnish detailed information on tube operation under varied conditions. There are 20 standard leads, 12 pair of thermocouple leads, and three high-frequency coaxial leads connected to a terminal board on one side of the chamber where measurements can be made, it was explained. Four pipes through the walls of the "tube icebox" also make provision for the testing of forced-air and water-cooled tube types.

Amplidyne Booster Inverter Developed for Railway Trains

Announcement of a new amplidyne booster inverter which will provide railway trains with an economical supply of 60-cycle a-c electric power ample for radios, movies, public-address systems, fluorescent lighting, and improved air conditioning has been made by the General Electric Company.

Mounted under the railway car, it changes the car's axle-driven generator or battery from direct to alternating current and gives constant voltage and frequency without excessive losses. The inverter is rated 6.25 kva, 5 kw, and 0.8 power factor to invert direct current of from 90 to 160 volts to 230-volt 3-phase 60-cycle current with voltage and frequency closely held by means of a separately mounted regulator. The amplidyne booster inverter consists of an inverted converter (synchronous converter running from the d-c side) with an amplidyne mounted on the same shaft, operating at approximately 1,800 rpm.

Aluminum Substitutes for Copper. The United States Rubber Company, New York, N. Y., recently announced that it will substitute aluminum for copper in some of its building wire and cable now going into production. This action was taken as a means of alleviating the wire shortage resulting from the temporary scarcity of copper, the company stated, and expressed the belief that it was removing an obstacle to building construction. The aluminum insulated wire will be made in all sizes, as approved by the Underwriters' Laboratories, with the same over-all diameter as equivalent copper sizes.

Sarnoff Honored for 40 Years in Radio. David Sarnoff (M '23) president of the Radio Corporation of America, ventured some predictions on future scientific advances at a dinner given recently in the Waldorf Astoria Hotel, New York, N. Y., to commemorate his 40 years of service to radio. Some of the advances foreseen by Doctor Sarnoff, who admitted that forecasting the developments of science is an almost impossible task, were control of the weather by man, delivery of mail by radio, portable communication sets that will enable one person to talk with another anywhere, transformation of climates by diversion of ocean currents, world-wide television, and use of atomic energy to combat disease. Officials of RCA presented to Doctor Sarnoff the wireless key he used as an operator on April 14, 1912, the day the Titanic sank. On duty at that time at John Wanamaker's, he remained at his key for three days, reporting the names of the 706 survivors as they came in. Other speakers at the dinner included Owen D. Young, retired chairman of the board of the General Electric Company, and Doctor Karl T. Compton (F '31) president of Massachusetts Institute of Technology.

JOINT ACTIVITIES

Engineering Foundation Elects Officers for 1946

Officers for 1946–47 were elected at the annual meeting of the Engineering Foundation with Doctor A. B. Kinzel, vice-president of Union Carbide and Carbon Research Laboratories, Inc., and of the Electro Metallurgical Company, re-elected president of the Foundation. Other officers elected were:

Vice-Chairman—Doctor L. W. Chubb (F '21) director of the Westinghouse Research Laboratories, East Pittsburgh, Pa.

Director—Doctor Edwin H. Colpitts (F '12) retired vice-president of Bell Telephone Laboratories, Inc., New York, N. Y.

Secretary—John H. R. Arms.

Doctor Kinzel will represent the Engineering Foundation on the executive board of the National Research Council.

The Research Procedure Committee will be headed by Doctor Chubb for the coming year. Other members of the committee will be B. A. Bakhmeteff, consulting engineer and professor of civil engineering, Columbia University, New York, N. Y.; J. F. D. Smith, acting president of Purdue University and dean of the engineering school, Lafayette, Ind.; and F. F. Colcord, vice-president and manager of metal sales for the United States Smelting, Refining, and Mining Company, New York, N. Y.

The executive committee for the year will consist of Doctor Kinzel, J. Schuyler Casey, R. H. Chambers, C. R. Jones (M '30), and O. E. Buckley (F '29). Mr. Arms is secretary of the committee.

The Engineering Foundation is one of the departments of the United Engineering Trustees, Inc., a corporation which was set up jointly by the four national engineering Founder Societies which have an aggregate membership of more than 88,000. The foundation was set up for "the furtherance of research in science and engineering, and the advancement in any other manner of the profession of engineering and the good of mankind." The four Founder Societies are the AIEE, the American Society of Civil Engineers, the American Institute of Mining and Metallurgical Engineers, and the American Society of Mechanical Engineers.

New ASA Director of Information. Frank MacMillen, until recently a member of the staff of the New York *Times* has been appointed director of information for the American Standards Association. He will supervise an expanded educational program designed to inform the public, as well as interested industrial and technical groups, concerning the activities of the association. Mr. MacMillen previously had been on the staffs of the Associated Press, the *Wall Street Journal*, and *Newsweek* magazine.

UET Elects 1946–47 Officers. J. P. H. Perry, vice-president of the Turner Construction Company, New York, N. Y., has been re-elected president of the United Engineering Trustees. Other officers elected are:

Vice-Presidents—Ralph M. Roosevelt, New Canaan, Conn.
William H. Harrison (F '31) vice-president of the American Telephone and Telegraph Company, New York, N. Y.

Treasurer—Albert Roberts of Minerals Separation North American Corporation, New York.

Assistant Treasurer—C. R. Jones (M '30) Eastern transportation manager, Westinghouse Electric Corporation, New York, N. Y.

Secretary—John H. R. Arms.

The United Engineering Trustees is a corporation set up jointly by the four Founder Societies for the advancement of the engineering arts and sciences in all their branches. It functions through two departments: the Engineering Foundation and the Engineering Societies Library.

LETTERS TO THE EDITOR

INSTITUTE members and subscribers are invited to contribute to these columns expressions of opinion dealing with published articles, technical papers, or other subjects of general professional interest. While endeavoring to publish as many letters as possible, Electrical Engineering reserves the right to publish them in whole or in part or to reject them entirely. Statements in letters are expressly understood to be made by the writers. Publication here in no wise constitutes endorsement or recognition by the AIEE. All letters submitted for publication should be typewritten, double-spaced, not carbon copies. Any illustrations should be submitted in duplicate, one copy an inked drawing without lettering, the other lettered. Captions should be supplied for all illustrations.

An Analysis of Electromagnetic Forces

To the Editor:

Reference is made to a letter from Alfred Gronner which appeared in the June 1946 issue of *Electrical Engineering*, pages 300–02, commenting on my article "An Analysis of Electromagnetic Forces" which was published in the October 1945 issue, pages 351–6. In his letter Mr. Gronner, borrowing from certain relativity concepts, offers a mathematical study in support of the popular belief that "parallel currents attract."

While this belief is not an inherent part of conventional electromagnetic theory, it is a natural result of the prevailing methods of interpreting the subject. Following this belief to its logical conclusion leads to the inconsistencies I spoke of in my article, and the combination of relativity and orthodoxy proposed in Mr. Gronner's letter does not resolve them. In fact, it appears to add to their complexity by presenting a derivation for the force between current streams which involves the velocity of the observer. This would make the activities of nature dependent upon the viewpoints of observers, a result opposed to all sense of logic, and one which I was able to eliminate in my analysis. Otherwise one is led to inquire how nature will act in the absence of observers. A situation such as that proposed by Mr. Gronner would shake the very foundations of electrostatics, and leave no basis for electrodynamics except in the mind of the observer. If this conclusion seems a little strong, it is inescapable if one accepts at its face value Mr. Gronner's statement: "F_{2y} is a force *due to the relative motion of the observer and the charges*." The italics are his, presumably intended to emphasize the conclusion just stated, following which he derives from the value of F_{2y} the formula for the electromagnetic force between current streams. I prefer to believe that electromagnetic forces are natural forces, having a real existence, and not merely existing in the mind of the observer. Nature must be presumed to proceed in an orderly fashion without the requirement of human intervention.

The approach to this subject as outlined in my article is receiving considerable attention by a number of investigators, particularly in the more scientific circles, while there is increasing evidence that even in engineering circles thoughtful individuals are becoming doubtful of the basis for certain prevailing beliefs. F. W. Warburton of the University of Kentucky, in letters to the editor of the *Physical Review*,[1,2] offers formulas for the forces between charges based solely on the relative velocity and acceleration of the charges. I might interpolate here that my article in the October 1945 *Electrical Engineering* was based on, and limited to, the concept of relative velocities, and showed that this concept led to the demonstrable electromagnetic forces; but that I have applied the method to relative acceleration also, with the result that these lead to an explanation of the effects of electromagnetic induction, such as induced voltages, self- and mutual-induction, radiation, and the like.

As to the conventional treatment of Maxwell's concept of the electromagnetic field, definite dissatisfaction is voiced by R. B. Lindsay and H. Margenau, of Brown University and Yale University respectively, who in a book[3] on physics state: "It failed, of course, to meet the demands of the flood of electrical phenomena discovered during the last few decades of the century, but in its very failure it has influenced profoundly the course of present-day physics." In view of such opinions I believe one is justified in questioning the validity of the conventional approach, even when employed by so eminent an authority as J. A. Stratton of the Massachusetts Institute of Technology (my own alma mater) who in his book[4] on the subject (one of the textbooks in the McGraw-Hill international series in physics) starts out in the first paragraph with an acceptance of Maxwell's equations and postulates them as the basis of his development of the subject.

Doubt is simmering in Great Britain also. Alfred O'Rahilly[5] of University College, Cork, has devoted the major portion of a large and masterful volume to this very subject. The terms "doubt" and "simmering" are, perhaps, a bit mild to use in referring to O'Rahilly's strongly stated opinions, but may seem better fitted to some of the comments appearing in the technical press. Reference is made, for instance, to articles by G. W. O. Howe[6,7] in the *Wireless Engineer*, London, and I. A. Robertson[8] in the *Philosophical Magazine*.

There is a substantial but rather scattered literature on the subject, the foregoing references illustrating some of the more recent only. Even such a limited group as this, however, demonstrates the chaotic situation existing with regard to the fundamental theory behind the matter. For instance, references 6 and 8 discuss the same point, which concerns an apparent paradox in the case of charges moving at right angles to each other. Both commentators point out that conventional analysis leads to the conclusion that there is a force on one particle due to the other, but no force on the second due to the first, a result at variance with Newton's third law of motion. Robertson attempts to explain away this paradox by drawing on the work of Ampere, and arrives at the conclusion that there is no force on either charge. Howe arrives at the same conclusion by an involved and unconvincing application of Maxwell's displacement current concept. The relative velocity concept followed in my article leads to the conclusion that there are equal forces on both charges. It is interesting to note in this connection that my formula (equation 4 in my article) is substantially that of Ampere, except that I use relative velocities and Ampere used absolute velocities, whatever they may mean today.

Howe's other article (reference 7) and Mr. Gronner's comments illustrate another divergence of opinion. In this case Howe attempts to explain the pinch effect as a purely electrostatic phenomenon, while according to Mr. Gronner it is an electromagnetic phenomenon by reason of the motion of the negative charges relative to the observer. But in my article I show that the relative velocity concept explains it as an electromagnetic phenomenon due to the motion of the negative charges relative to the positive charges.

These instances illustrate only some of the chaotic conditions and the shades of opinion existing in presumably well-informed circles. Something should be done to clear up the situation, and that is what I am trying to do, if only to the extent of stirring up the subject.

Echoes of these more technical aspects of the subject have appeared to some extent in the pages of *Electrical Engineering* other than in my article. In the October 1945 issue, page 381, the same issue in which my article appeared, B. Litman of the Westinghouse Electric Corporation, in a letter to the editor, raises a question on the very point at issue between Mr. Gronner and myself. I have often wondered whether Mr. Litman read the article and was satisfied with my explanation. Also in *Electrical Engineering* there have appeared, from time to time, essays for recreation. These have been interesting but many of them would have had no point if there were not certain widely held misconceptions upon which they were able to build. Other articles have appeared in the past, notably those of many years ago by Carl Hering, which I read at the time with much interest and which I am glad to acknowledge had much to do with arousing my interest in the subject.

In the analysis of electromagnetic forces as developed in my article, I arrive at a concept for the effects of relative velocity, which I embody in two propositions. It is presumably these two propositions to which Mr. Gronner refers when he accuses me of postulating new laws of nature. Actually the concept is not new at all, having been formulated by Wilhelm Weber almost exactly 100 years ago. It was developed further by a number of the leading scientists of the latter half of the 19th century, including Clausius and Riemann, and even to some extent by H. A. Lorentz, until obscured by the ideas of Maxwell. Now the wheel turns again. The Maxwellian approach is found wanting, and the older concept is receiving renewed attention. My two propositions, which I developed strictly in accordance with this concept, lead directly to my formula for the force between moving charges. This is the formula mentioned above which, except for using relative velocities instead of absolute velocities, is equivalent to the one Ampere developed as far back as 1823. It proves to be a powerful method, and with our greater knowledge of the nature of matter and electricity, it is hoped that we can forge it into a more complete theory of electromagnetism than did our scientific predecessors.

To continue with a specific consideration of Mr. Gronner's comments, it seems superfluous to point out that the action of current flow must produce forces of a definite and predictable magnitude. Take the case of short-circuit currents, which can produce destructive effects if precautions are not taken to provide adequate support for the conductors carrying these currents. Certainly the dangerous effects possible in such cases cannot be dependent upon the viewpoint of an observer. But application of Mr. Gronner's method and assignment of various velocities to his observer certainly will lead to a wide variety of inconsistent results, as will be obvious. It does not seem necessary to give specific numerical examples.

The trouble with Mr. Gronner's analysis is that he has considered only a special case (that of equal parallel velocities) and attempts to build up a general theory of electromagnetism from it. It is suggested that his method be tried for the general case, in which the velocities of the charges may have any assignable individual values. It then will be found that the analysis cannot be carried out without eliminating the observer and considering the velocity of the charges relative to each other. This case will bring out also the main mathematical error in Mr. Gronner's analysis, which occurs in his equation 18 wherein v^2 is equated to $v_1 \times v_2$. This error is obscured in Mr. Gronner's work, as he does not use the subscripts 1 and 2, but he definitely uses these two velocities as the individual velocities of the two current streams in equations 20 and 21. In his special case they have equal magnitudes but are none the less separate mathematical entities and cannot be derived from v^2. This will be clear in the general case wherein v_1 does not equal v_2, and, therefore, $v_1 \times v_2$ cannot be equal to v^2. In this case there is no means of making the transition from v^2 to $v_1 \times v_2$, and I should like to see Mr. Gronner's method of accomplishing this step.

I might mention, for the benefit of anyone interested in analyzing Mr. Gronner's letter, that there are a number of typographical errors which make the mathematics somewhat difficult to follow. These appear to be truly typographical in nature, and no criticism can be made of Mr. Gronner for their occurrence. On the other hand, I do not believe he has added to the clarity of his case by including the appendix for the derivation of the integral of equation 16. The integrand in equation 16 is a standard form of a quadratic to the minus 3/2 power and is given in any short table of integrals. It yields the result obtained if one makes the approximation that $x/\sqrt{x^2+\beta^2a^2} = 1$, when $x = \infty$. This is just as accurate an approximation as tanh $\alpha = 1$, when $\alpha = \infty$.

With regard to Mr. Gronner's introduction of the factor $\sqrt{1-v^2/c^2}$, this appears to be an *ad hoc* assumption without good authority. Admittedly, it is used freely in relativity analysis in many ways, but not to my knowledge as a factor to relate forces in different frames of reference as he has used it. It would appear that Mr. Gronner has introduced considerable novelty at this point. Moreover, if such a term is to be used at all to relate observed forces, I believe the manner of using it as proposed by Mr. Gronner will lead to results at variance with the law of conservation of momentum. This will be seen from the following analysis.

Conservation of momentum requires that what an observer notes as force must be proportional to the product of what he calls mass and the rate of change of velocity, regardless of the frame of reference. Since Mr. Gronner's problem is concerned only with components in the *y* direction, it is necessary to consider only the *y* components of force and acceleration. Taking the problem as he has set it up, and using his nomenclature, an observer in the stationary frame of reference will note that the force is

$$F_y = md^2y/dt^2 \qquad (1)$$

while an observer in the moving frame of reference will note the force as

$$F_y{}^* = m^*d^2y^*/dt^{*2} \qquad (2)$$

The problem is to find out whether Mr. Gronner's analysis yields this invariant relationship. That is, will an observer in the stationary system agree that equation 2 is correct for an observer in the moving system? Following Mr. Gronner

$\beta = \sqrt{1-v^2/c^2}$

$F_y = \beta F_y{}^*$

$y = y^*$ and therefore $d^2y = d^2y^*$

$t^* = (t - vx/c)/\beta$, and therefore $dt^{*2} = dt^2/\beta^2$ and $dt^2 = \beta^2 dt^{*2}$, a relationship which is mathematically sound and has the proper operational effect in the term d^2y/dt^2

It also commonly is accepted in relativity circles that

$$m^* = m/\beta, \text{ or } m = \beta m^*$$

Substituting these values in equation 1 gives

$$\beta F^* = \beta m^* d^2y^*/\beta^2 dt^{*2}$$

$$\text{or } F^* = m^* d^2y^*/\beta^2 dt^{*2} \qquad (3)$$

It is clear that equation 3 does not agree with equation 2, and the observer in one system will not agree with the observer in the other system. It would appear that if the factor $\sqrt{1-v^2/c^2}$ is to be used at all, Mr. Gronner's equation for the relationship between F_y and $F_y{}^*$ should be $F_y = F_y{}^*/\beta$ instead of $F_y = \beta F^*$. But if this should be the case, the term $(1-v^2/c^2)$ in his equation 16 would vanish, and the subsequent value of F_{3y} would be zero, as predicted in my article. It thus is seen that, if Mr. Gronner's result is accepted, one must discard the principle of the conservation of momentum, whereas if this principle is to be retained, one cannot accept Mr. Gronner's result.

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Sign of Reactive Power

To the Editor:

The article, "Sign of Reactive Power," which appeared in the November 1946 issue of *Electrical Engineering*, pages 512–16, is a most excellent presentation of the considerations which led a special subcommittee of the AIEE Standards committee to recommend that the present internationally recognized convention for the algebraic sign of reactive power should be reversed. The presentation is accurate and shows conclusively that, although either convention is possible mathematically, the present standard is the one which yields a vector power diagram which is similar to the admittance diagram and which is therefore the more useful for handling problems involving many individual loads connected in parallel. In spite of this result the subcommittee chose the opposite convention, because they believed that otherwise they logically would be compelled to abandon the even more useful concept that an "inductive load absorbs (positive) reactive power."

I believe, however, that their statement "no workable compromise which would allow the standard to remain unaltered and at the same time would meet the objections to it could be found" is unduly pessimistic. It seems to me that there is an alternative which will enable us to "eat our cake and have it, too," and which will cause the resulting minus sign to appear at a place where it will not do serious violence to recognized mathematical conventions.

The problem is to reconcile two desiderata, each of which is based on a long-recognized and significant fact, but between which there exists a certain implicit incompatibility. The first is

A. An overwhelming majority of the a-c circuits dealt with in modern electrical engineering are of the constant-voltage parallel-load type, and hence it is desirable to analyze them by vector diagrams for power which are similar to the vector diagrams of admittance and of current.

It was this consideration which led the AIEE committee on this subject in 1932 and the International Electrotechnical Commission in 1935 to select the present sign for reactive power. The second and equally significant fact is

B. An overwhelming majority of loads happen to be inductive, and it is therefore desirable that in metering and load dispatching the terminology should be consistent with the flow of a positive quantity analogous to energy from an overexcited generator to an induction motor.

This concept, that there is something which flows into an inductive load while reactive power is being maintained at its terminals, has become of great practical value. Var-hour meters have been invented to meter this imaginary quantity, and varmeters have been invented to indicate its rate of flow. Yet hitherto there has been no name for the quantity itself. This lack of adequate nomenclature has led to the unfortunate practice of speaking about "flow of reactive power," "absorption of reactive power," or for short "flow of reactive" and even "flow of x." Of course, strictly speaking reactive power does *not* flow but merely exists as a measure of the rate of flow of something else, just as active power does not flow but exists as a measure of the rate of flow of energy. Yet it is this concept, that something flows and that whatever does flow should be defined as positive, combined with the false notion that it is the reactive power which really flows, which has led to the recent request that the sign of reactive power be reversed.

I would like to propose as an alternative the following three-point program which seems to me to yield a more satisfactory solution than that offered by the subcommittee.

Point 1 is to coin a word to serve as a name for the quantity which is thought of as flowing into an inductive circuit and of which the var-hour is the unit. I propose the word, "quadergy" derived, of course, by combining syllables of "quadrature" and of "energy" (the term, "reactive energy" should not be used but should be reserved for the actual energy which alternately is stored and returned by an inductive circuit. This latter quantity is true energy, though it seldom is realized how small it is. For instance a circuit at which the reactive power is maintained at 1,000 kilovars stores and returns an amount of energy, the maximum value of which is only 0.0007 kilowatt-hours). The proposed definition is:

Quadergy 05.21.051

The quadergy which has been supplied by a source to a load during a time interval is the integral with respect to time of the reactive power at the points of entry of the source, taken over the time interval.

The value of quadergy is given in var-hours when the reactive power is in vars and the time is in hours.

Note: In describing the operation of an a-c system with reactive loads, it has been found convenient to assume the existence of this quantity, which is analogous to energy, which is inherently positive, and which is assumed to flow out of a properly excited source into an inductive load.

C. It follows from definition 05.21.050 that the quadergy supplied to an inductive load is positive because the reactive power at the points of entry of a source of quadergy is positive, while that at a sink of quadergy is negative.

Point 2 is to set up a standard convention for marking the scales and terminals of varmeters and related apparatus, so that there may be uniformity of practice among different power companies, and so that the instruments will indicate conveniently the flow of quadergy in the way most desired by load dispatchers. Presumably this will involve marking based on an "in and out" flow of quadergy rather than on "plus or minus" signs of reactive power or on "lag or lead" of current.

At its meeting in the spring of 1946 the AIEE committee on instruments and measurements authorized the formation of a subcommittee for this purpose.

Point 3 is to leave the present definition of the sign of reactive power unchanged.

It will be seen that Point 2 satisfies the desideratum *B*, and Point 3 satisfies the desideratum *A*, and that Point 1 merely gives recognition to and makes more convenient the use of a concept which long experience has shown to be of great value in the power industry.

With this procedure the incompatibility will show up in a difference in the practice of connecting varmeters and wattmeters. This difference can be expressed in either of two ways.

1. If we still retain the arbitrary convention that an upscale deflection of an instrument is *always* to correspond to a positive value of the quantity (active power or reactive power) indicated by the instrument, then it appears that at a given set of points of entry, which divide a complete circuit into two portions, the wattmeter indicates the active power of one portion while the varmeter indicates the reactive power of the other portion. The magnitude of either power is necessarily the same for both portions of the circuit.

2. The alternative and simpler way of describing this situation is to say that we scrap the arbitrary convention which connects the direction of an instrument deflection and the sign of the indicated quantity

Viewed on either alternative this situation is closely parallel to, and no worse than, that of a center-zero ammeter in a storage battery circuit. Here custom requires that an upscale deflection correspond to a condition of charge. If the ammeter is in the positive lead of the battery, this upscale deflection does correspond to a positive current, but, if the ammeter happens to be connected in the negative lead, either (1) the ammeter must be considered as measuring the current in the charging circuit instead of the current in the battery or else (2) the upscale deflection on charge must be taken to correspond to a negative current, that is, a flow of electric charge out of the load.

If the actual connection of wattmeters and varmeters on the switchboard were made directly on the basis of fundamental definitions, this situation might be objectionable. In practice, however, the electrician wires the instrument in accordance with the manufacturer's blueprint. The blueprint is based on current practice for wattmeters and on the standard as set up in Point 2 for varmeters, and I believe no difficulties will be created. In fact, at present the generator wattmeters are connected to read upscale for energy flow *out* of the generator, while motor wattmeters are connected to read upscale for energy flow *into* the motor. I propose that we similarly connect generator varmeters to read upscale when quadergy is flowing *out* of the generator and connect motor varmeters so as to read upscale when quadergy is flowing *into* the motor. This ignores the fact that at the generator the active power is negative while the reactive power is positive, because I believe that this inconsistency is much less disturbing than an abandonment of desideratum *A*.

Quadergy is of course a highly artificial and intangible concept, yet it would seem that definite progress should result if engineers should give to this "airy nothing, a local habitation and a name."

F. B. SILSBEE (F '42)

(Chief, division of electricity, National Bureau of Standards, Washington, D. C.)

NEW BOOKS....

"Heating Ventilating Air Conditioning Guide 1946." This 24th edition of the guide contains an enlarged technical data section and a new grouping of subjects "to provide a more logical progression from fundamental principles to specific application." A new chapter on "Fluid Flow" has been added, and important additions and revisions have been made in the chapters: Thermodynamics, Air Contaminants, Instruments and Measurements, Heat Transmission Coefficients of Building Materials, Physiological Principles, Air Conditioning in the Prevention and Treatment of Disease, Cooling Load, Heating Boilers and Furnaces, Gravity Warm Air Systems, Mechanical Warm Air Systems, Steam Heating Systems and Piping, Pipe Insulation, Dehumidification by Sorbent Materials, Refrigeration, Air Distribution, Air Duct Design, and Owning and Operating Costs. The chapter on terminology has been revised to obtain better conformity with the terms of basic science rather than the less accurate forms often adopted by specialists. Besides the 51-chapter technical data section, a manufacturers' catalogue data section, and the roll of membership of the American Society of Heating and Ventilating Engineers are included in the volume. Complete indexes to the technical and catalogue sections are provided. The technical data section is divided into eight sections: princspals, human reaction to atmospheric environment, heating and cooling loads, combustion and consumption of fuel, heating systems and equipment, air conditioning, special applications, installation and testing codes. The American Society for Heating and Ventilating Engineers, New York 10, N. Y., 1946, $6^{1}/_{4}$ by $9^{1}/_{4}$ inches, 1,279 pages plus the roll of membership, cloth, $6.

The following new books are among those recently received at the Engineering Societies Library. Unless otherwise specified, books listed have been presented by the publishers. The Institute assumes no responsibility for statements made in the following summaries, information for which is taken from the prefaces of the books in question.

ASTM STANDARDS ON RUBBER PRODUCTS. Prepared by ASTM Committee D-11 on Rubber Products; Methods of Testing, Specifications, March, 1946. American Society for Testing Materials, Philadelphia 2, Pa. 540 pages, illustrated, 9 by 6 inches, paper, $3.25 (to ASTM members, $2.20). More than 75 specifications and tests covering natural and synthetic rubbers are collected in this publication. The largest section covers general methods of analysis, identification, and testing. The following sections deal with electric tests; insulated wire and cable; hose, belting, gloves, tape, and so forth, nonrigid plastics; and tests and specifications for miscellaneous products not previously covered. There is a section on nomenclature and definitions, and a number of proposed specifications are included to solicit comment.

CIRCUIT ANALYSIS. By C. E. Skroder and M. S. Helm. Prentice-Hall, Inc., New York, N. Y., 1946. 288 pages, illustrated, $9^{1}/_{4}$ by 6 inches, cloth, $5.35. The object of this book is to provide in a single volume a wide selection of laboratory circuit experiments, including a very comprehensive coverage of the related circuit theory and laws, the theory and limitations of the necessary instruments, and the methods of making measurements. The experiments are presented as problems to induce the student to adopt an analytical approach in determining the experimental data needed, the circuits to be used, the measurements to be made, and the interpretation of the observed and calculated data for a rational solution.

CURRENTS IN AERIALS AND HIGH-FREQUENCY NETWORKS. By F. B. Pidduck. Clarendon Press, Oxford, England; Oxford University Press, New York, N. Y., 1946. 97 pages, illustrated, 9 by $5^{1}/_{2}$ inches, cloth, $2.50. This book is an account of previously unpublished investigations on currents in aerials, based on a little-known paper of Pocklington. Fundamental formulas are deduced and extended to soldered networks. The theory of transmission lines is established as that of a system of two aerials, the untuned Y feeder is considered in detail, and a theory of aerails parallel to the earth is worked out. The necessary tables for calculations are included.

INDUCTANCE CALCULATIONS. By F. W. Grover. D. Van Nostrand Company, New York, N. Y., 1946. 286 pages, illustrated, $9^{1}/_{4}$ by 6 inches, cloth, $5.75. This volume has been prepared with the idea of providing for each special type of inductor a single simple formula that will involve only the parameters that naturally enter together with numerical factors obtainable from tables. General principles and working methods are given in the introductory chapters. The separate types of circuits are covered in the succeeding chapters, including the necessary tables with the last three chapters devoted to auxiliary table of functions and special formulas.

PLASTICS FOR ELECTRICAL AND RADIO ENGINEERS. By W. J. Tucker and R. S. Roberts. Technical Press, Ltd., Gloucester Road, Kingston Hill, Surrey, England, 1946. 148 pages, illustrated, 9 by $5^{1}/_{4}$ inches, cloth, 12*s*. This handbook provides the electronic engineer with essential data relating to the application of plastics in the electrical and radio industry. Molding and manufacturing procedure and the advantages and limitations of plastic materials are discussed in a full and practical manner. Insulation and testing problems are dealt with in detail, and tables giving physical properties are presented in a manner useful to the electronic industry.

CALCULUS. By F. H. Miller. Second edition. John Wiley and Sons, New York, N. Y.; Chapman and Hall, London, England, 1946. 416 pages, illustrated, $8^{1}/_{2}$ by $5^{1}/_{2}$ inches, cloth, $3.50. This book is designed to give the student a comprehension of the basic concepts and methods of calculus, presenting the subject both as an important branch of mathematics and as a tool for practical use. Additions to the new edition include a discussion of graphical differentiation, a summary of the processes of integration, a discussion of approximate integration, and numberous formulas and theorems. Answers are given to the odd-numbered exercises only, the balance of the answers being available spearately to teachers.

PIEZOELECTRICITY. By W. G. Cady. McGraw-Hill Book Company, New York, N. Y., 1946. 806 pages, illustrated, 9 by $5^{3}/_{4}$ inches, cloth, $9. The entire field of piezoelectricity is covered in this comprehensive treatise, including related areas of elasticity, dielectrics, optics, and magnetism. Crystallography and the general properties of crystals lead up to the discussion of quartz and Rochelle salt with their special applications. A unified account is given of experimental results, with many formulas, numerical data, and an extensive bibliography.

LES TENSEURS EN MÉCANIQUE ET EN ELASTICITÉ. By L. Brillouin. Dover Publications, New York, N. Y., 1946. 364 pages, illustrated, $9^{1}/_{2}$ by 6 inches, cloth, $3.75. This book presents an exposition of tensor analysis and its applications in theoretical physics. The range of the book is indicated by the chapter headings: vector geometry; pseudo-tensors; the principal differential operators; parallelism, covariant, and contravariant derivative; Riemann space; use of Riemannian geometries in analytical mechanics; transition of wave mechanics; elasticity and waves in elastic solids; theory of solids and quantum mechanics. There is an index.

LES RADIATIONS. By C. Fabry. Librairie Armand Colin, Paris, France, 1946. 220 pages, illustrated, $6^{1}/_{4}$ by $4^{1}/_{2}$ inches, paper, 60 francs. The fundamental properties of radiations, their sources, and effects are discussed in this small volume. The author presents a critical study of recent developments in producing, analyzing, and measuring thermal and optical radiations, including fluorescence and phosphorescence. The final chapter describes biologic and therapeutic applications and uses in analytic chemistry and photochemistry.

LABOR PROBLEMS. By W. V. Owen. Ronald Press Company, New York, N. Y., 1946. 570 pages, illustrated, $9^{1}/_{4}$ by 6 inches, cloth, $4.50. Intended to present a survey of the principles and problems of labor economics from a broad viewpoint, this book covers the subject in four main sections: the economic characteristics of labor and the nature of the labor market; the labor market in operation, including supply and demand, wages, hours, working conditions, and the mobility of labor; the control of the labor market exerted by the trade unions, the employer, and the state; and security in the labor market as affected by old age, unemployment, and time lost through strikes and disagreements.

PSYCHOLOGY IN INDUSTRY. By N. R. F. Maier. Houghton Mifflin Company, Boston, Mass., New York, N. Y., and Chicago, Ill., 1946. 463 pages, illustrated, $8^{1}/_{4}$ by $5^{1}/_{4}$ inches, cloth, $3. The author discusses in a nontechnical but systematic manner the various aspects of the industrial situation to which the principles of psychology are applicable. Beginning with the fundamental problem of why people act as they do, he covers morale and motivations, the allowance for variations in individuals, the use of psychological tests and the analysis of skills, fatigue conditions and the prevention of accidents, and psychological factors in labor turnover. The book is designed for a wide audience from the student to top management.

RADAR. By J. F. Rider and G. C. B. Rowe. John F. Rider, Publisher, Inc., New York 16, N. Y., 1946. 72 pages, illustrated, 11 by $8^{1}/_{2}$ inches, paper, $1. The underlying principles of radar are explained in simple language with a description of the basic radar set. The use of radar during the war by land, sea, and air forces is described, including countermeasures. A brief survey of the future of radar is added in conclusion.

STEAM POWER PLANT AUXILIARIES AND ACCESSORIES. By T. Croft, editor, revised by D. J. Duffin, 2nd edition. McGraw-Hill Book Company, Inc., New York, N. Y., and London, England, 1946. 583 pages, illustrated, $8^1/_4$ by $5^1/_2$ inches. cloth, \$5. A practical manual for the operating engineer, this book has been revised to conform with the changes that have taken place in the 24 years since it was published originally. Topics on which considerable new material has been added include reciprocating and centrifugal pumps, methods of boiler feeding, feedwater heaters, economizers and air preheaters, condensers, steam piping of power plants, and steam traps. As before, there are review questions and problems, and a technical data section has been added.

ELECTRON MICROSCOPE. By E. F. Burton and W. H. Kohl. Second edition. Reinhold Publishing Corporation, New York, N. Y., 1946. 325 pages, illustrated, $9^1/_4$ by 6 inches, cloth, \$4. The opening chapters cover the principles of the optical microscope, including a discussion of wave motion as a means of propagation of energy. The author proceeds to a description of electromagnetics and the electron, and then presents the physical principles upon which the operation of the electron microscope is based. Both the electrostatic and magnetic types are covered. Important practical applications of the electron microscope are considered in the last chapter. The book is well illustrated with sketches and microphotographs, and there is a 20-page bibliography.

GRUNDZÜGE DER THEORETISCHEN LOGIK. By D. Hilbert and W. Ackermann. Second enlarged edition. Dover Publications, New York, N. Y., 1946. 133 pages, tables, $8^3/_4$ by $5^1/_2$ inches, cloth, \$2.50. An introduction to the subject of symbolic logic, this book demonstrates the application of strict formulas and methods for their manipulation to the field of philosophical logic. Four chapters deal respectively with: the propositional calculus; the calculus of classes; the simple function calculus, that is, the calculus of classes of individuals and of multiple relations between individuals; and the logical paradoxes and the use of a theory of type to avoid paradox.

MACRAE'S BLUE BOOK. Fifty-third annual edition. MacRae's Blue Book Company, 18 East Huron Street, Chicago, Ill., 1946. 3,736 pages, illustrated, 11 by 8 inches, cloth, \$15. This annual directory gives a complete listing of manufacturers in the United States, classified according to products. The first 400 pages of the volume contain an alphabetical list of manufacturers, producers and wholesalers, with capital ratings, location of branches, and, in some cases, addresses of local distributors. The third and last section of the volume is an alphabetical list of trade names.

ELECTRIC DISCHARGE LAMPS. By H. Cotton. Chapman and Hall, London, England, 1946. 435 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, 36*s*. The structure of atoms and molecules is explained in sufficient detail for an understanding of the fundamental principles involved in light production in discharge lamps. Descriptions are given of the various kinds of lamps in present use, with their construction and operating characteristics. The control of electric discharge lamps is discussed, and a chapter is devoted to the nature of fluorescence and the applications of fluorescent materials to these lamps.

ENGLISH–FRENCH AND FRENCH–ENGLISH TECHNICAL DICTIONARY. By F. Cusset. Chemical Publishing Company, Brooklyn, N. Y., 1946. 590 pages, $6^3/_4$ by 5 inches, cloth, \$5. Giving both English to French and French to English translations, this dictionary, as shown on the title page, covers metallurgy, mining, electricity, chemistry mechanics, and science. Phrases as well as words are given and in many cases appear directly under each of the important words. A few basic conversion tables appear at the end of the book.

LABOR–MANAGEMENT ECONOMICS. By W. V. Owen in collaboration with Stevenson, Jordan and Harrison, Inc., Management Engineers. Ronald Press Company, New York, N. Y., 1946. 121 pages, diagrams, $8^1/_4$ by $5^1/_4$ inches, cloth, \$2. Part I, on employer–employee economics, deals with management, production, risks, and costs on the employer's side of the question and with wages, labor economics, and unionism, as more directly related to the employees' interests. It ends with a chapter on industrial relations in general. Part II, the economic framework, is concerned with some of the economic forces, concepts, and relations that provide the economic mechanics for producing, marketing, and consuming goods.

ELECTRONICS IN INDUSTRY. By G. M. Chute. McGraw-Hill Book Company, New York, N. Y., and London, England, 1946. 461 pages, illustrated, $8^1/_2$ by $5^1/_4$ inches, cloth, \$5. To give a broad introduction to the use of electronic circuits and equipment is the purpose of this book. It outlines the industrial uses of tube circuits and gives detailed explanation of a large number of electronic equipments now serving in industrial plants. No previous knowledge of tubes is assumed, the early chapters being devoted to the necessary fundamentals, and, since the book is intended for users of equipment already available, no design information is presented.

ATOMIC SPECTRA. By R. C. Johnson. Methuen and Company, Ltd., London, England. Distributed by Chemical Publishing Company, Brooklyn, N. Y., 1946. 120 pages, illustrated, $6^3/_4$ by 4 inches, cloth, \$2. Following a brief introduction on the production of spectra, the book discusses spectra of the hydrogen type, one electron revolving around a positive nucleus, and spectra of atoms with two or more electrons. The structure of atoms is considered, and two chapters deal with the effect of special conditions and applied fields on spectral lines. The organization of the periodic table on the basis of the electronic structure of the elements is demonstrated, and spectroscopic instruments and procedures are described.

PRINCIPLES OF PHYSICS. Volume 2, Electricity and Magnetism. By F. W. Sears. Addison-Wesley Press, Cambridge 42, Mass., 1946. 434 pages, illustrated, 9 by 6 inches, cloth, \$5. A college textbook, this volume deals with the elements of electricity and magnetism on the basis of a thorough preliminary background of mathematics and general physics. Separate chapters are devoted to ferromagnetism, electromotive force, and the basic principles of electronics.

FUNDAMENTALS OF ALTERNATING-CURRENT MACHINES. By A. Pen-Tung Sah. McGraw-Hill Book Company, New York, N. Y., and London, England, 1946. 466 pages, illustrated, 9 by $5^3/_4$ inches, cloth, \$5. Written from the viewpoint of the operating man rather than that of the design engineer, this book lays particular stress on the operating problems of electric machines. The student is shown not only how to test his equipment, but also what tests to make in order to obtain quickly and accurately the requisite parameters from which he may compute the performance. Thus design formulas are avoided, while the experimental determination of the machine parameters is given full prominence after the theory is explained. Only those theories and experimental methods capable of being generalized are treated, and for directness the theory is developed from the standpoint of circuits.

WAVE PROPAGATION IN PERIODIC STRUCTURES. By L. Brillouin. McGraw-Hill Book Company, Inc., New York, N. Y., and London, England, 1946. 247 pages, illustrated, $8^1/_2$ by $5^1/_4$ inches, cloth, \$4. Based on lectures given at the University of Wisconsin, this volume includes a variety of problems having a common mathematical background. They extend from electrical engineering to electromagnetism and wave mechanics of the spinning electron. The book includes explanations of electric filters, rest rays, anomalous optical reflection, selective reflection of X rays or electrons from a crystal, and omission of energy dissipation.

UNDERSTANDING MICROWAVES. By V. J. Young. John F. Rider Publisher, New York 16, N. Y., 1946. 385 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, \$6. This book explains the fundamental problems encountered in the field of ultrahigh frequency research and production and how they are surmounted—problems met in the design and operation of wave guides and coaxial lines; resonant cavities as they function in the magnetron, the dynatron, and the klystron; the theory and design of antennas used in conjunction with the transmission and reception of microwaves. The underlying electromagnetic and electrostatic theories are interpreted in the early chapters. Section II contains a comprehensive collection of terms, ideas, and theorems with thorough definitions and explanations.

TRIGONOMETRY. By H. K. Hughes and G. T. Miller. Second edition. John Wiley and Sons, New York, N. Y.; Chapman and Hall, London, England, 1946. 175 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, \$2.50. Intended for use in a first course in trigonometry for college freshmen, this book aims particularly at clearness and ease of understanding. To this end, certain changes in the order and amount of material have been made in the new edition. More than customary attention is given to the law of cosines and to the use of logarithms with this law. No tables are included in the volume.

PHYSICAL AND CHEMICAL EXAMINATION OF PAINTS, VARNISHES, LACQUERS, AND COLORS. By H. A. Gardner and G. G. Sward, Tenth edition. Henry A. Gardner Laboratory, 4723 Elm Street, Bethesda, Md., 1946. 652 pages, illustrated, 12 by $10^1/_4$ inches, fabrikoid, \$18.50. The most comprehensive work in its field, this revised edition presents current standard test methods and experimental methods, as well as some that may temporarily be obsolete. It includes charts of color families and also charts of colors used by the Army and the Navy. The book covers all types of examination used in its field. It is well illustrated and has many charts and diagrams.

ORGANIZED LABOR AND PRODUCTION. By M. L. Cooke and P. Murray. Revised edition. Harper and Brothers, New York, N. Y., and London, England, 1946. 277 pages, illustrated, $8^1/_4$ by 6 inches, cloth, \$2.50. A prominent expert in scientific management and the head of a leading union join in setting forth what they both believe is a basis for greater collaboration among employers and organized workers to assure greater production under fair terms of employment. The book covers the responsibilities and activities of both labor and management and discusses some of the questionable practices of both.

METALLIC CORROSION, PASSIVITY AND PROTECTION. By U. R. Evans. Appendix by A. B. Winterbottom. Longmans, Green and Company, New York, N. Y., 1946. Revised edition, 863 pages, illustrated, $9^1/_2$ by $5^3/_4$ inches, cloth, \$14. The material in the new edition has been rearranged with considerable addition of recent developments. The first two chapters present simple examples of corrosion and passivity and a study of thin films. Corrosion under varying conditions, such as with and without oxygen, is described in succeeding chapters, followed by a discussion of the influence of stress, structure, crevices, and the like. Four chapters are devoted to protection by various methods, and the book ends with a section on testing. Each chapter is divided into three parts: the scientific basis; the technical aspects including practical problems; and the mathematical, or quantitative, treatment.

PAMPHLETS • • •

The following recently issued pamphlets may be of interest to readers of "Electrical Engineering." All inquiries should be addressed to the issuers.

How to Use Carbon Tetrachloride Safely: Tips to the Foreman. Safety Research Institute, Inc., 420 Lexington Avenue, New York 17, N. Y., single copies, five cents. Quantity prices will be quoted.

The Action of Embeco in Concrete and Mortars (second edition). The Master Builders Company, 7016 Euclid Avenue, Cleveland 3, Ohio, 34 pages, no charge.

Celanese Synthetics for the Electrical Industry, Celanese Plastics Corporation, 180 Madison Avenue, New York 16, N. Y., 20 pages, no charge.

TRANSACTIONS SECTION

Preprint of Corresponding Pages From the Current Annual AIEE Transactions Volume

An Electronic Frequency Meter and Speed Regulator

ELLIS LEVIN
ASSOCIATE AIEE

AN electronic frequency meter and speed regulator has been developed to measure and hold the speed of a small steam turbine within plus or minus one-half of one per cent of any speed between 10,000 and 100,000 rpm.

This equipment fills a need for precise measuring and speed control apparatus required in the development and testing of very high speed machinery. In addition, the new circuits are of such basic nature that they readily can serve other applications for which no suitable commercial equipment is available. The frequency meter in its present form can be used as a general purpose laboratory instrument for the precise measurement of frequency or the accurate indication of small deviations from a given frequency. It gives an indication which is stable to about one or two rpm under normal line voltage variations, and its accuracy can be made to approach one-tenth of one per cent.

For use on machinery operating at speeds up to 100,000 rpm, the measuring instrument must not unbalance (and in some cases must not load) the moving part; therefore a photoelectric speed pickup is employed which requires no mechanical connection to the turbine shaft. The speed pickup, which can be placed several inches away from the turbine, merely "looks at" alternate black and polished areas on the surface of the turbine shaft and produces an electric signal whose frequency is some known multiple of the rotational frequency of the shaft.

Then in place of counting the revolutions per minute of the shaft we have the problem of measuring the frequency of an electric signal. The frequency meter used for this purpose must meet the following requirements:

1. Accurate within plus or minus one-half of one per cent or better.
2. Continuously adjustable over a frequency range of at least ten to one.
3. Capable of operating from an input signal having a very distorted wave form such as would be provided by a photoelectric or magnetic pickup.
4. Capable of providing an error signal proportional to deviation from any desired frequency to operate a servomechanism for regulating the turbine speed.

None of the commercial frequency meters are able to meet all of these requirements. A new frequency measuring circuit therefore has been devised which is essentially an electronic frequency meter of the recurrent capacitor discharge type but modified to operate as a "null-balance" circuit instead of a direct-reading device. By this simple modification the accuracy is increased from plus or minus two per cent to the order of one-tenth of one per cent and means are provided for obtaining a d-c or an a-c error voltage proportional to variation of shaft speed from the desired value. This voltage is used for indicating speed variations and for automatic speed regulation. It also operates a protective relay which will stop the machine in the event of an excessive speed deviation.

General Description

The desired speed is set up on a 3-dial decade speed selector having increments of 10,000, 1,000, and 100 rpm. The difference between selected speed and measured speed is called speed error. Its magnitude and algebraic sign are indicated on a large speed error meter, which has two ranges: plus or minus 250 rpm and plus or minus 1,000 rpm. A small speed meter, 0 to 100,000 rpm, is provided to show the approximate speed at which the turbine is running.

An electropneumatic servomechanism actuated by a speed error signal operates a steam control valve to regulate the steam pressure at the input to the turbine and hold the turbine speed at the desired value. The servomechanism is provided with a proportional air pressure "follow-up" and a manual reset control called the air pressure control dial. The steam control valve also is provided with proportional pressure follow-up so that the valve reproduces output pressure rather than valve position. In addition, the speed error signal contains a certain amount of error "anticipation" to minimize the response time of the system. By holding down a "bat handle" on the control panel the operator at any time can disconnect the speed error signal and assume manual control of the turbine speed, which he regulates by means of the air pressure control dial.

A block diagram of the complete speed control system is shown in Figure 1. The general scheme of operation is as follows: The turbine operator sets the speed selector dial to the number of revolutions per minute at which he desires the turbine to run. By means of the air pressure control dial he manually brings the turbine up to speed while watching the approximate turbine speed indicated by the speed meter. When the turbine speed comes within 4,000 or 1,000 rpm (operator's choice) of the desired preset value, the system switches into automatic control and the air pressure control dial becomes a fine speed adjustment. The operator trims the air pressure to make the speed error meter stabilize at zero. The frequency meter and servomechanism then hold the speed within plus or minus one-half of one per cent of the desired value. When a major load change is made, the operator retrims the air pressure control dial to return the speed error to zero, and the system regulates about the new control

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The author wishes to acknowledge the contributions of Doctor C. K. Stedman, for his suggestion of the frequency measuring circuit; G. H. Stoner, for joint design of the complete frequency meter and regulator; and other persons who assisted in the work.

point. To measure a given speed at which the turbine is running, the operator adjusts the speed selector until the speed error meter read zero, and then reads the setting of the speed selector dial.

If for any reason the actual turbine speed goes either above or below the selected speed by more than 1,000 or 4,000 rpm (operator's choice), automatic protection circuits will shut off the steam supply and stop the turbine. In addition, the turbine automatically will be shut down by failure of any of the electric circuits. The speed error meter and automatic protection circuits function regardless of whether the system is in automatic or manual control.

The operator's control panel is shown in Figure 2. At the top is the frequency meter chassis with the speed selector in the center. On the left is the speed meter and an indicator light which shows when sufficient speed signal is present. On the right is the speed error meter, its range switch, and a light which indicates when the speed error is less than 1,000 or 4,000 rpm. Below the frequency meter is a panel containing steam and air pressure gauges. The next lower chassis contains the servo and automatic protection circuits, with the "bat handle" in the center and the air pressure control dial on the right. At the very bottom is the power supply chassis, with switch, fuses, and a monitor voltmeter.

At the present time, the turbine is being driven by high pressure air instead of steam, involving no changes, however, in the speed control equipment.

The Basic Frequency Measuring Circuit

The basic null-balance frequency measuring circuit is shown in Figure 3. It consists of a pair of push-pull connected triode switching tubes (V_1 and V_2) capacitor coupled to a diode bridge rectifier. The negative end of the bridge output resister R_B is grounded to the negative end of the common triode cathode resistor R_K and the negative side of battery B_1.

A square wave voltage, derived by the use of suitable shaping or trigger circuits from the signal whose frequency is to be measured is applied to the input terminals *A* and *B* in such a way that during one half-cycle the potential of terminal *B* is sufficiently negative to drive tube V_1 to cutoff. During the period of cutoff no voltage drop caused by tube current occurs in the resistor R_{L1}, and the capacitor C_1 will become charged to approximately the potential of battery B_1. The charging path is from ground, through battery B_1, resistor R_{L1}, capacitor C_1, diode tube V_4, rectifier output resistor R_B, and back to ground.

During the next half-cycle the signal applied to terminal *B* is within the operating range of tube V_1, and terminal *A* is sufficiently negative to drive tube V_2 to cutoff, so that tube V_1 will conduct and allow the potential of its anode to drop by an amount IR_{L1}, where *I* is the steady state plate current of the tube. The potential across the capacitor C_1 will drop by the same amount IR_{L1}, the discharge path being to ground through tube V_1 and resistor R_K in parallel with resistor R_{L1} and thence back to the capacitor C_1 through battery B_2 and diode V_3.

Since the electronic circuits are symmetrical, a similar charging and discharging of capacitor C_2 produces identical current pulses at alternate half-cycle intervals of the driving signal voltage, with the result that the charging current for the capacitors C_1 and C_2 flows through the resistor R_B in the manner illustrated in Figure 3a, and the discharging current of the capacitors C_1 and C_2 flowing through the resistor R_K takes the form shown in Figure 3b. The resistor R_B, therefore, carries only the transient charging current of capacitors C_1 and C_2, while the resistor R_K carries the steady current of the tubes V_1 and V_2 in addition to part of the transient discharge current of the capacitors C_1 and C_2.

In order to utilize this circuit as a frequency meter, it is only necessary to adjust discharge resistor R_K so that the average voltage developed across this resistor equals the average voltage across the rectifier output resistor R_B. When this has been done the voltmeter *V* will indicate zero.

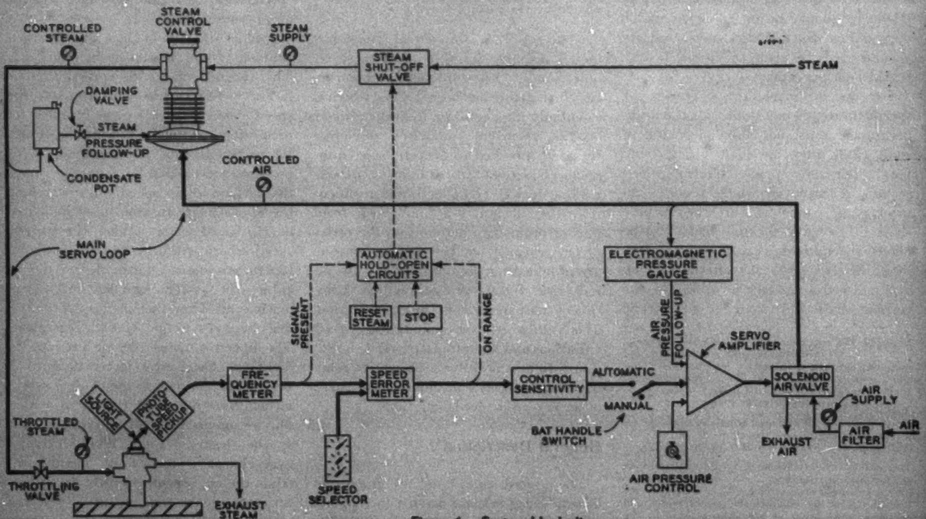

CONTROLLED STEAM
STEAM CONTROL VALVE
STEAM SUPPLY
STEAM SHUT-OFF VALVE
STEAM
DAMPING VALVE
STEAM PRESSURE FOLLOW-UP
CONDENSATE POT
CONTROLLED AIR
MAIN SERVO LOOP
AUTOMATIC HOLD-OPEN CIRCUITS
RESET STEAM
STOP
SIGNAL PRESENT
ON RANGE
ELECTROMAGNETIC PRESSURE GAUGE
AIR PRESSURE FOLLOW-UP
SERVO AMPLIFIER
FRE-QUENCY METER
SPEED ERROR METER
CONTROL SENSITIVITY
AUTOMATIC
MANUAL
BAT HANDLE SWITCH
SOLENOID AIR VALVE
AIR SUPPLY
AIR FILTER
AIR
EXHAUST AIR
AIR PRESSURE CONTROL
LIGHT SOURCE
PHOTO-TUBE SPEED PICKUP
THROTTLED STEAM
THROTTLING VALVE
TURBINE
EXHAUST STEAM
SPEED SELECTOR

Figure 1. System block diagram

Figure 2. The control cabinet

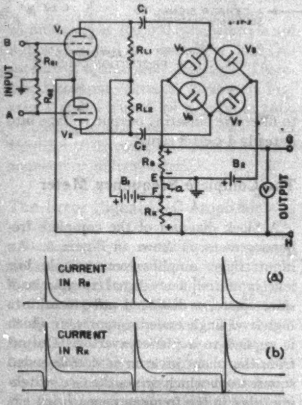


Figure 3. The basic null-balance frequency measuring circuit

If the frequency of the driving voltage applied to terminals *A* and *B* is raised, the heights and durations of the peaks in Figures 3a and 3b will not change, but will move closer together. The average current through resistor R_B is thus proportional to the frequency. However, the average current in resistor R_K is practically independent of frequency because the area of the discharge current peaks is only a fraction of the total area under the curve shown in Figure 3b. Consequently, the value of the resistor R_K required to balance the voltmeter *V* to zero will be very nearly proportional to the frequency of the signal applied to the input terminals *A* and *B*.

The equation of balance now will be derived. Since the circuit is symmetrical, either of the identical wire-wound resistors R_{L1} or R_{L2} can be represented by the single symbol *R*, and either of the high-stability mica capacitors C_1 and C_2 can be represented by the symbol *C*. R_p indicates the anode resistance, *I* the steady state anode current of either of the tubes V_1 and V_2, *f* the frequency of the signal voltage applied to the terminals *A* and *B*, R_B the value of the rectifier output resistor, and R_K the value of the discharge resistor required to balance the voltmeter *V* to zero. Each time a capacitor *C* is charged its voltage rises by an amount *IR*, and its charge rises by an amount *CIR*. The number of such charging pulses passing through R_B per second is 2*f*. Thus the average voltage across R_B is

$$2fCIRR_B \tag{1}$$

Similarly, it can be shown that the average voltage across R_K is

$$R_K I+\left[\left(\frac{IR_K R^2}{R+R_p+R_K}\right)C\times 2f\right] \tag{2}$$

If R_K is adjusted until the voltmeter *V* reads zero, then these two average voltages are equal, and

$$2fCIRR_B=R_K I+\left[\left(\frac{IR_K R^2}{R+R_p+R_K}\right)C\times 2f\right] \tag{3}$$

Solving for *f*,

$$f=\frac{R_K}{2RC\left[R_B-\dfrac{RR_K}{R+R_p+R_K}\right]} \tag{4}$$

Since the expression $RR_K/(R+R_p+R_K)$ in the denominator varies with R_K from zero to a very small percentage of the resistance of R_B, the calibration curve of *f* as a function of R_K is very nearly a straight line, or

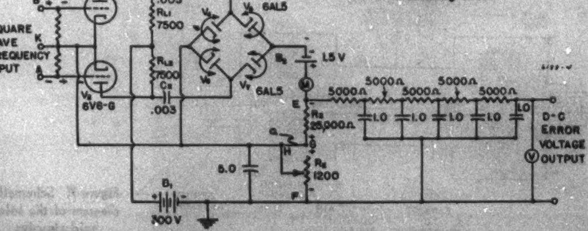
$$f\backsim\frac{R_K}{2RCR_B} \tag{5}$$

If the circuit is to be used as a frequency meter, it is only necessary to vary the discharge resistor R_K until the voltmeter *V* reads zero, and then compute the input signal frequency by means of equation 4 or 5, or obtain the frequency value directly from the calibration curve previously prepared by comparison with a frequency standard.

Such a circuit can be used as a speed regulator control if discharge resistor R_K is adjusted so that the voltmeter *V* reads zero when a pickup device on the shaft rotating at the desired speed is supplying substantially square wave voltage impulses via a suitable trigger circuit to the signal input terminals *A* and *B*. Variations in speed of the shaft then will produce a pulsating d-c error voltage at the output terminals *G* and *H* which can be smoothed by a suitable electric filter and impressed upon a servo amplifier for speed control purposes.

Equation 5 shows that the only components which materially affect the calibration are the capacitor *C* and the resistors R_K, R_B, and *R* (R_{L1} and R_{L2} in Figure 3). If these components are stable and have low temperature-coefficients, the performance of the circuit as a frequency meter or as a speed regulator will be accurate and stable. The stability of the circuit also is increased by the fact that the voltages across resistors R_B and R_K are both proportional to the steady state anode current *I* so that variations caused by tube characteristics and battery voltages cancel out. The battery B_2 reduces to a low value the current which initial velocities of electrons in the diodes would cause to flow through the diodes and R_B. Changes in the contact potentials of the tubes resulting from changes

Figure 4. The single-ended-output frequency measuring circuit



in heater voltage are small compared with the charging potential of the battery B_1. Equation 4 shows that moderate variations in the anode resistance R_p of the tubes V_1 and V_2 have only a minor effect on the frequency calibration.

The Single-Ended Frequency Measuring Circuit

The circuit shown in Figure 3 and discussed thus far utilizes a double-ended input signal balanced to ground and produces an error voltage such that neither output terminal is grounded. A single-ended input can be utilized, if desired, by eliminating half of the symmetrical capacitor charging and discharging circuit, namely, V_2, R_{L2}, C_2, V_6, V_7, and R_{O2}. The output circuit can be modified to produce an error voltage with one output terminal grounded, by interchanging the wire *a* (Figure 3) with the voltmeter V, so that terminals G and H are joined together when the voltmeter is connected between terminals E and F.

The turbine speed control system described in this paper employs the frequency measuring circuit arranged for a double-ended input and grounded output, as shown in Figure 4. In this modified circuit the voltmeter V is connected between terminals E and F, and the terminals G and H are connected together. This joins the two measuring resistors R_B and R_K at their positive voltage ends rather than at their negative ends (as was done in Figure 3). In the circuit of Figure 4 at the bridge output, resistor R_B will conduct capacitor discharging current instead of charging current, but the circuit performance will be substantially the same as in the basic circuit which is shown in Figure 3.

Also in Figure 4, a milliammeter M has been inserted to measure the current in R_B. As this current will be proportional to the input frequency but subject to errors caused by variations in line voltage and tube characteristics, the meter M is used only for an approximate indication of the input frequency or speed. But the value of R_K required to balance the voltmeter V to zero at a given input frequency will be independent of variations in line voltage and tube characteristics to an accuracy approaching one-tenth of one per cent. The setting of R_K which gives zero output error voltage is used as a precise measure of the input frequency. The capacitor connected across the cathode resistor R_K, and the five RC sections connected to output terminals E and F serve to filter the pulsating output voltage into a steady d-c error signal.

Figure 5. Block diagram of the complete frequency meter

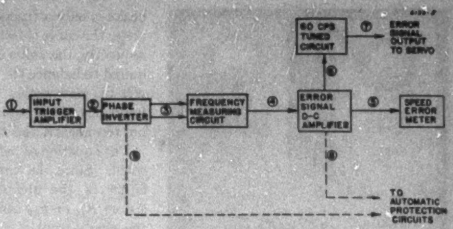
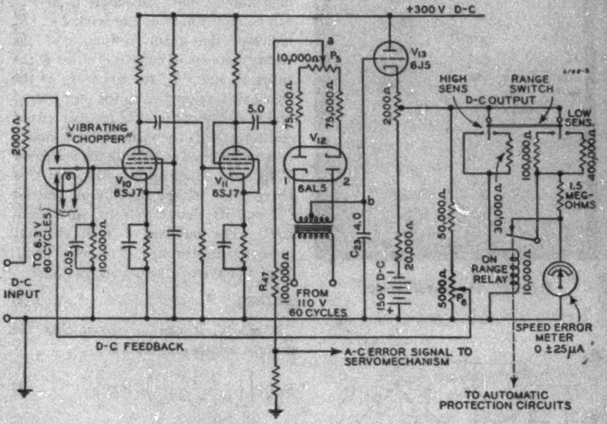

①
INPUT TRIGGER AMPLIFIER
②
PHASE INVERTER
③
FREQUENCY MEASURING CIRCUIT
④
ERROR SIGNAL D-C AMPLIFIER
⑤
SPEED ERROR METER
⑥
60 CPS TUNED CIRCUIT
⑦
ERROR SIGNAL OUTPUT TO SERVO
⑧
⑨
TO AUTOMATIC PROTECTION CIRCUITS

Figure 6 (below). The error signal d-c amplifier circuit


+300 V D-C
2000 Ω
VIBRATING "CHOPPER"
TO 6.3 V 60 CYCLES
D-C INPUT
V_{10} 6SJ7
V_{11} 6SJ7
0.05
100,000 Ω
5.0
a
10,000 Ω
P_5
75,000 Ω
75,000 Ω
V_{12}
6AL5
1
2
b
C_{23} 4.0
R_{47} 100,000 Ω
FROM 110 V 60 CYCLES
V_{13} 6J5
2000 Ω
20,000 Ω
150 V D-C
HIGH SENS
RANGE SWITCH
D-C OUTPUT
LOW SENS.
100,000 Ω
400,000 Ω
1.5 MEG-OHMS
50,000 Ω
30,000 Ω
5000 Ω
P_6
ON RANGE RELAY
10,000 Ω
SPEED ERROR METER 0 ±25μA
D-C FEEDBACK
A-C ERROR SIGNAL TO SERVOMECHANISM
TO AUTOMATIC PROTECTION CIRCUITS

The Complete Frequency Meter

A block diagram of the complete frequency meter is shown in Figure 5. An input trigger amplifier converts the low level input frequency signal (this can be of sine, square, or distorted wave form) into high level single-ended square wave which is applied to a phase inverter. Output from the phase inverter is double-ended square wave which drives the two switching tubes in the frequency measuring circuit.

Full scale d-c error signal from the frequency measuring circuit is about 25 volts for 100,000-rpm speed error or 0.25 millivolts per revolution per minute deviation from the selected speed. In order to achieve a measuring and control accuracy of plus or minus one-half of one per cent if any speed between 10,000 and 100,000 rpm the circuit should be sufficiently sensitive to indicate changes of at least 0.1 to 0.05 per cent. For 0.1 per cent speed deviation the d-c output error voltage is 2.5 millivolts at 10,000 rpm and

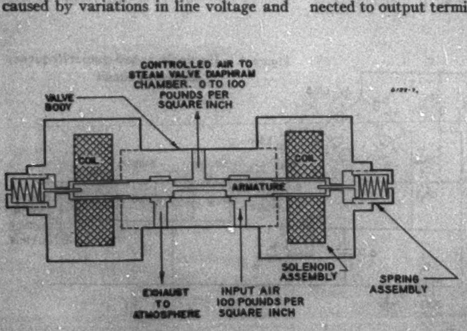

CONTROLLED AIR TO STEAM VALVE DIAPHRAM CHAMBER. 0 TO 100 POUNDS PER SQUARE INCH
VALVE BODY
COIL
COIL
ARMATURE
SOLENOID ASSEMBLY
SPRING ASSEMBLY
EXHAUST TO ATMOSPHERE
INPUT AIR 100 POUNDS PER SQUARE INCH

Figure 7. Schematic diagram of the solenoid air valve

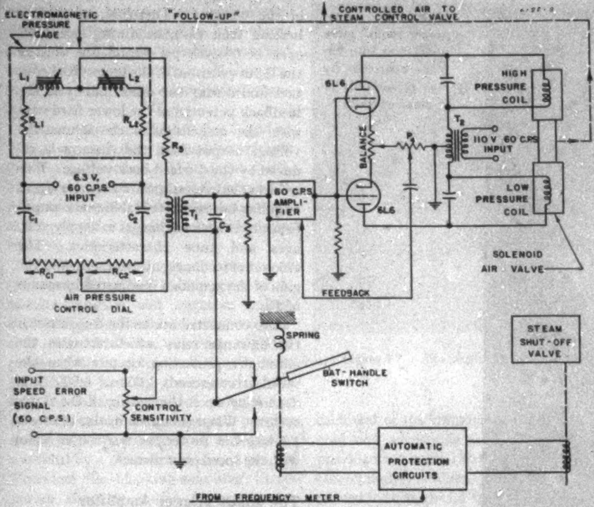

ELECTROMAGNETIC PRESSURE GAGE
"FOLLOW-UP"
CONTROLLED AIR TO STEAM CONTROL VALVE
L1
L2
RL1
RL2
R3
6.3 V, 60 CPS INPUT
C1
C2
T1
C3
RC1
RC2
AIR PRESSURE CONTROL DIAL
60 CPS AMPLIFIER
6L6
6L6
BALANCE
P1
T2
110 V 60 CPS INPUT
HIGH PRESSURE COIL
LOW PRESSURE COIL
SOLENOID AIR VALVE
FEEDBACK
SPRING
BAT-HANDLE SWITCH
STEAM SHUT-OFF VALVE
INPUT SPEED ERROR SIGNAL (60 CPS.)
CONTROL SENSITIVITY
AUTOMATIC PROTECTION CIRCUITS
FROM FREQUENCY METER

Figure 8. The servo circuit

25 millivolts at 100,000 rpm, which is not sufficient signal to operate a high resistance d-c voltmeter directly. Therefore the output voltmeter *V* shown in Figures 3 and 4 should be a vacuum tube voltmeter, which generally employs an amplifier to drive a panel type d-c microammeter or milliammeter.

The Error Signal D-C Amplifier

Because of the instabilities associated with conventional direct-coupled amplifiers a special stabilized, sensitive, discriminating d-c amplifier was devised for use in this frequency meter. This amplifier utilizes the established technique of "chopping" the d-c signal into square wave alternating current and passing it through a more easily stabilized a-c amplifier; but in addition it employs a familiar phase-discriminating rectifier to change the alternating current back to high level direct current, and incorporates d-c negative feedback to produce a high degree of over-all stability.

The circuit of the error signal d-c amplifier is shown in Figure 6.

The input d-c "chopper" is effectively a single-pole double-throw mechanical switch which opens and closes continuously at the rate of 60 cycles per second. It consists of a vibrating reed contactor driven between two stationary contacts by means of a 6.3-volt 60-cycle-per-second exciter coil. As the reed oscillates between the two fixed contacts it assumes the potential of the upper contact for approximately one-half of the 1/60-second cycle and the potential of the lower contact during the other half of the cycle. Therefore, when a direct voltage is applied between the two fixed contacts it will be converted to a 60-cycle-per-second square wave voltage across the vibrator output impedance connected between the reed and the lower fixed contact. The output alternating voltage will be in phase with the 110-volt 60-cycle-per-second line when the applied direct voltage is of a given polarity (say positive) and 180 degrees out of phase with the line when the input direct current is of the opposite polarity (negative).

Consider for the present that the lower fixed contact is grounded. Then the amplitude of the 60-cycle-per-second square wave from the vibrator will be approximately equal to the d-c error voltage and proportional to speed error. The phase will reverse when the speed error goes through zero. Connected across the 100,000-ohm vibrator output resistor is a 0.05-microfarad capacitor which serves to by-pass the high frequency transient peaks which occur during the "make" and "break" switching intervals.

The low level a-c error voltage is amplified in the conventional 6SJ7 stages V_{10} and V_{11}, and then impressed upon the phase-discriminating half-wave rectifier V_{12}, which operates as follows: Via the injection transformer T_1, the 110 volt a-c line voltage switches the two diodes of V_{12} on and off at the rate of 60 cycles per second and in synchronism with the 60-cycle-per-second vibrating reed in the d-c "chopper" at the input to the error signal amplifier.

The 60-cycle-per-second speed error voltage applied between point *a* and ground cannot reach point *b* when the diodes are cutoff; but during the half-cycle of line voltage that the diodes conduct there is a path of finite resistance from *a* to *b* allowing pulsating direct cur-

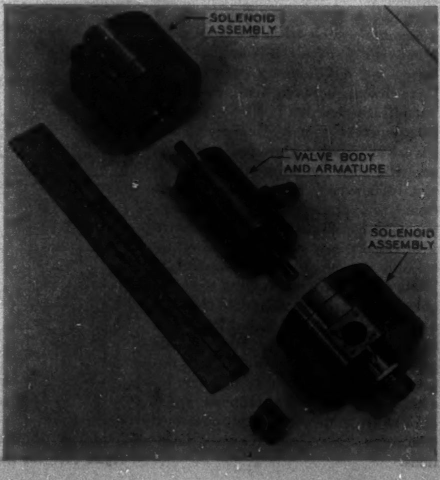

SOLENOID ASSEMBLY
VALVE BODY AND ARMATURE
SOLENOID ASSEMBLY

Figure 9. The solenoid air valve partly disassembled

THROTTLING VALVE
TURBINE
CONTROL VALVE

Figure 10. Temporary control valve installation with turbine operated by high pressure air instead of steam

rent to be fed to capacitor C_{23} through two parallel paths, each consisting of one-half of P_5, one $75K$ resistor, one diode, and one-half of the secondary winding of T_1. (The diodes are balanced so that the switching voltage from T_1 causes no potential difference between points *a* and *b*.) Because it comes from the d-c chopper, the a-c signal applied to point *a* is either in phase or 180 degrees out of phase with the 110-volt 60-cycle-per-second line. If *a* becomes positive when the diode plate 2 becomes positive the pulsating direct current will flow from *a* to *b* and through C_{23} to ground, producing across C_{23} an output voltage such that point *b* is positive with respect to ground. Similarly if point *a* is driven negative when diode plate 2 is negative, then the pulsating direct current will flow from ground through C_{23} and from *b* to *a*, producing across C_{23} an output voltage such that point *b* is negative with respect to ground. Capacitor C_{23} smooths the pulsating direct current to give a steady output voltage of the same polarity but greater magnitude than the d-c speed error signal applied to the d-c chopper at the input to the amplifier. The output high level d-c error voltage is impressed upon the cathode follower V_{13} which drives the zero-centered speed error meter to the right or left of zero to indicate positive or negative speed deviations.

By means of a voltage divider connected across the cathode follower output terminals, a portion of the d-c output voltage is fed back to the lower fixed contact of the input "chopper." The circuit is phased, by means of T_1, so that the potential at *h* is of the same polarity as the input potential at the upper fixed contact of the vibrator. Therefore, as the vibrating reed assumes during one half-cycle of 60-cycle-per-second line voltage the input potential of the upper fixed contact and during the other half-cycle the feedback potential of the lower fixed contact, the magnitude of the alternating voltage output from the vibrator is reduced by the d-c feed-back voltage. This reduces the over-all gain of the error signal amplifier but serves to stabilize the amplifier output against changes in supply voltages and tube characteristics. The amount of feedback, and consequently the gain of the amplifier, is adjusted by means of P_6.

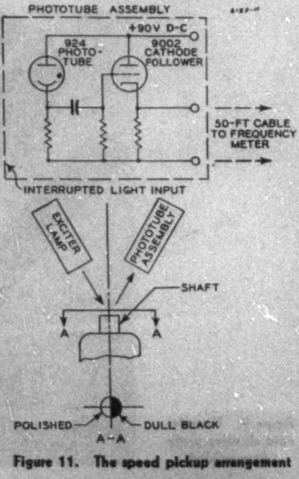


Figure 11. The speed pickup arrangement

Also connected across the d-c output is the on-range relay which actuates the automatic protection circuits when the speed error exceeds 1,000 or 4,000 rpm, depending upon the setting of the range switch. When energized it also inserts a 1.5-megohm protective resistor in series with the speed error meter.

The Input Trigger Amplifier

There are various shaping amplifiers which can be used to operate the frequency measuring circuit. The one selected for this frequency meter is a new reliable "go–no go" circuit. That is, when the input speed signal amplitude exceeds a critical minimum voltage of about 0.15 volt (at any frequency from zero to at least 5 kc), the amplifier will trigger and provide high level square wave output of constant amplitude and one-half the input frequency. If the input signal amplitude falls below the critical value, the circuit will give no output at all. This feature is utilized by the automatic protection circuits to stop the turbine and summon the operator in the event of erratic or insufficient input signal from the photoelectric speed pickup.

The circuit diagram is shown in Figure 13. The input tube V_0 is an amplifier to feed the required 10,000-ohm input to the first trigger tube V_1. With the plate, cathode, and grid resistance values as shown, V_1 can be set up to have two stable operating points. The bias and feed-back resistors are adjusted until the tube is operating in one equilibrium condition and almost at the point where it will change to the second equilibrium condition. Then a small input signal will trigger the tube from one operating condition to the other. Output can be taken between ground and either the plate or the number 2 grid.

Although the tube will be triggered by either a rising or decreasing wave front as soon as the amplitude crosses a given critical value, better square wave results when the triggering impulse is a sharp pulse, in which case a small capacitor connected between the plate and the number 1 grid adds to the reliability of operation.

The output from the number 2 grid of V_1 has squared sides but not a perfectly flat top and bottom. This signal is differentiated in the output *R-C* branch (25 micromicrofarads, 10,000 ohms) to obtain sharp pulses for tripping a second trigger tube V_3.

The differentiated output appears at the number 1 grid of V_2 as a series of alternate positive and negative pulses. But the time interval between a given positive pulse and the adjacent negative pulse might be different in length from the interval between the negative pulse and the following positive pulse if the speed input signal has unequal positive and negative half cycles such as might be provided by a photoelectric speed pickup. Therefore the blanking amplifier V_2 removes alternate pulses (by successive positive and negative blanking), producing at its second cathode a new series of equidistant pulses that are positive only. These are differentiated in the output *R-C* branch (10 micromicrofarads, 500,000 ohms) to obtain very sharp double-ended pulses which are used to fire the second trigger tube V_3. A given pulse triggers the tube in one direction and the following pulse trips it in the opposite direction. The resultant output from V_3 is a good square wave voltage having equal positive and negative half cycles. Its frequency is exactly half that of the input speed signal. This single ended square wave is fed to a 6N7-G phase inverter which produces double-ended square wave to drive the two switching tubes V_1 and V_2 in the frequency measuring circuit.

The Servomechanism

As shown in Figure 1, the diaphragm-actuated steam control valve is operated by an electropneumatic servomechanism, which receives 60-cycle-per-second speed error signal from the frequency meter unit. To supply regulated air to the steam control valve, at 0 to 100 pounds per square inch, the servomechanism utilizes a solenoid air valve shown schematically in Figure 7.

The air valve contains a balanced armature that slides in a cylinder which is provided with an annular metering port on each end and a central "controlled pressure" outlet connected to the diaphragm chamber of the steam control valve. At each end of the cylinder housing, a solenoid and adjustable spring assembly is provided to position the armature within the cylinder. When the armature is moved to the right far enough to open the inlet port, high pressure air will flow around the undercut middle section of the armature and through the center outlet to the diaphragm of the steam control valve. The pressure built up in the diaphragm chamber will depend upon how much and how long the inlet port is open. When the armature is returned to the neutral position so that both ports are closed, the output pressure will remain at the built-up value until the armature is moved to the left to open the exhaust port or until the air leaks out around the armature bearing surfaces. Thus by proper positioning of the armature, the output pressure can be set to any value between 0 and 100 pounds per square inch.

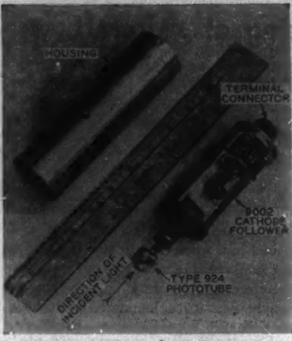


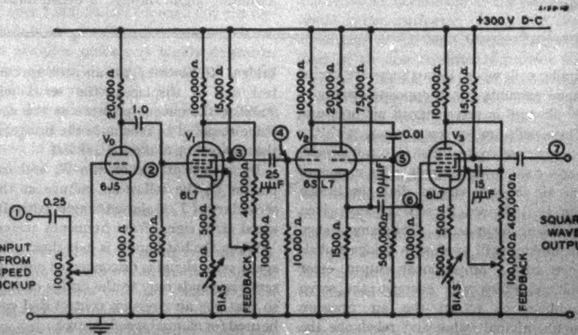
Figure 12. The phototube assembly

Each of the solenoid coils is connected as the plate load of a 6L6 power output tube in a phase-discriminating servo amplifier shown in Figure 8. The two 6L6 plate circuits are energized, 180 degrees out of phase, by the 60-cycle-per-second line via the high voltage transformer T_2 whose secondary is grounded at its center-tap. At zero a-c input to the 6L6 grids, the tubes are balanced to supply equal plate currents to the two solenoid coils. A 60-cycle-per-second error signal applied to the grids such that it is in phase with the plate voltage of the upper tube will be out of phase with the plate voltage of the lower tube; therefore the plate current of the upper tube will increase and that of the lower tube will decrease. This will result in a differential magnetic pull on the valve armature, which will move the armature toward the inlet port to increase the output air pressure. Similarly, if the phase of the input error signal is reversed, the lower tube will conduct more plate current than the upper one, moving the armature toward the exhaust port to decrease the output air pressure.

The current in each solenoid is half-wave rectified 60-cycle-per-second, partly smoothed by means of a 0.5-microfarad shunt capacitor. The remaining a-c component of coil current is utilized to buzz the armature at 60 cycles per second and minimize static friction in the valve bearings. A portion of the second-harmonic cathode voltage is fed back to the input voltage amplifier to provide added stability and an over-all gain control.

The 60-cycle-per-second sine-wave input to the servo amplifier consists of speed error signal (coming in through a "control sensitivity" attenuator and the bat handle switch) in series with air pressure follow-up signal (from the pressure gauge bridge circuit via transformer T_1) The electromagnetic pressure gauge consists of two

Figure 13 The input trigger amplifier



Electrical Anti-icing of Aircraft Windshields

J. D. McCRUMM
ASSOCIATE AIEE

PREVENTION of ice formation on the aircraft windshield presents difficult problems. In addition to optical quality, the thermal and structural characteristics of suitable transparent materials, such as glass and certain plastics, must be given careful consideration. Further, an important design criterion for an aircraft windshield is its ability to withstand impact loads such as those which result when a large bird strikes the windshield. Contemporary windshield design practice usually calls for a laminated structure, commonly consisting of a layer of polyvinyl butyral, encased between two plates of tempered glass. The impact resistance of such a construction has been shown to be dependent upon the vinyl temperature, the optimum temperature being of the order of 110 degrees Fahrenheit. Direct correlation thus exists between impact resistance and anti-icing ability in windshield design. The regulations of the Civil Aeronautics Board specify that the windshield and supporting structure of a transport airplane shall be designed to resist the impact of a 4-pound bird when the airplane is flying at sea level cruising speed.[1] It is apparent, therefore, that it may be necessary to supply heat to a laminated windshield continuously, regardless of the prevalence of icing conditions, and further that the amount of heat must be controlled within the limits specified by impact resistance as well as deicing.

Five schemes or combinations thereof have been advanced for the prevention of ice formation on the windshield:

1. Fluid and wiper.
2. Hot air.
3. Resistance wires embedded in the windshield.
4. Radiant heat (infrared).
5. A transparent electrically-conductive coating.

It is the purpose of this paper to examine the basic considerations for the design of a windshield anti-icing system, and to discuss in some detail the pertinent characteristics peculiar to each scheme. The design will be considered primarily as a windshield ice-prevention problem, with impact resistance as the major secondary consideration.

Heat Requirements

Thermal energy is removed in several ways from a heated windshield which is being flown in an icing environment. Assuming flight through a dense formation of super-cooled water droplets, effective anti-icing will be achieved by maintaining the outer surface of the windshield at a temperature above freezing, say 35 degrees Fahrenheit. The heat absorbed from the surface can be grouped into four basic categories:

1. The loss by forced convection and radiation.
2. The heat of liquid required to raise the temperature of the water droplets striking the surface.
3. The loss by evaporation of that portion of water required to maintain 100 per cent vapor saturation in the surface boundary layer.
4. The edge losses by conduction to the structure.

A fifth category, kinetic heating,* should be considered, representing a gain of heat.

The relative importance of kinetic heating is dependent upon the amount of free water in the cloud. For a low water density and small droplets, the droplets may be evaporated entirely before they strike the windshield surface. For a higher water concentration, water will strike the surface and at once be vaporized so that the surface remains dry; in this case, the kinetic heat equals or exceeds the heat losses. Insufficient kinetic heating is characterized by a sharp reduction in the temperature of the surface (with a consequent icing over) unless auxiliary heat is supplied. Complete wetting of the surface will be brought about by a relatively high concentration of water, and all five factors plus auxiliary heat achieve importance in an anti-icing heat balance analysis.

The relative magnitude of the thermal losses for a given airplane and windshield are dependent upon the outside air temperature, water droplet size, amount of water per unit volume, altitude, and aircraft velocity. Two other determining factors, the surface boundary layer velocity distribution and amount of water catch, are influenced by those structural conditions which determine the aerodynamic flow characteristics over the windshield. Briefly consider each previously mentioned heat loss.

Loss by forced convection in wet air is greater than in dry air because of the higher specific heat of wet air; the radiation loss is relatively small. The heat of liquid required to raise the water temperature is dependent upon the amount of water catch on the windshield and the temperature differential through which it is raised. The loss by heat of evaporation

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* In wet air Prandtl's number approaches unity, and kinetic heating becomes proportional to the airplane velocity squared.

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The author wishes to acknowledge the assistance of the air conditioning group and research laboratories of the Douglas Aircraft Company in the described investigation and the co-operation of the Pittsburgh Plate Glass Company in supplying design data. Appreciation is expressed for permission to reproduce from reference 2 the data shown in Figure 1.

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similar coils whose inductances are varied equal amounts but in opposite directions by changes in servo output air pressure. These coils are connected along with the air pressure control dial into a 60-cycle-per-second bridge circuit. By means of the air pressure control dial, the bridge can be balanced to zero at any given operating air pressure. Then any change in pressure will produce a bridge unbalance and a proportional output error voltage, which will energize the servo amplifier to return the air pressure to its original value and rebalance the bridge. If, however, the air pressure control dial is displaced, the servo will stabilize the output pressure at the new value required to rebalance the bridge at the new setting of the control dial.

By means of the resistor R_4 and capacitor C_3 the follow-up voltage at the secondary of T_1 is phased exactly with the speed error signal from frequency meter.

When the bat handle is held down, the speed error signal is disconnected and the servo responds only to the bridge circuit, so that the air pressure control dial can be used for manual speed control.

requires more detailed explanation. As the temperature of the air in the boundary layer over the windshield increases, because of the acquisition of auxiliary heat, it possesses a greater ability to absorb water. If the air at the initial temperature is considered 100 per cent saturated, a lower degree of saturation will exist at the higher temperature. Because of the resulting differential vapor pressure conditions, liquid will be evaporated from the water droplets and the wet windshield surface, and a considerable amount of thermal energy will be extracted in the process. Loss by conduction at the frame is a function of the frame configuration and thermal contact at the glass edge.

The temperature specifications for maximum impact strength have been mentioned. Figure 1, taken from reference 2, indicates a maximum strength when the vinyl plastic is maintained at 110 degrees Fahrenheit, but little deviation from the maximum strength results over a temperature range of from 100 to 120 degrees Fahrenheit. When the vinyl is at a temperature corresponding to that of cabin air, say 70 degrees Fahrenheit, the impact strength based on penetration velocity is reduced by about 45 per cent. This is a serious diminution. The data of Figure 1 refer to a specific windshield and frame design. Other similar designs will have different absolute maximum impact resistance, but the maximum will occur at about the same vinyl temperature.

It is evident that if impact resistance be considered a design criterion of an anti-icing windshield, the thermal requirements will be influenced accordingly. This is particularly true when the time schedule of operation is considered, since adequate impact resistance may require application of heat continuously while the airplane is in flight; anti-icing may be required only at infrequent intervals. The relative heat demand is modified by variations in glass and vinyl thickness, method of heat application, and plane at which heat is developed initially, as well as differences in the flight atmosphere.

Design Considerations

Definite factors influence the adoption and design of a windshield anti-icing scheme:

1. Anti-icing ability.

2. Impact resistance. Proper structural design is involved, possibly supplemented by heat.

3. Weight.

4. Optical properties. Freedom from annoying reflections, adequate light transmission, freedom from optical distortion, and so forth, are necessary.

5. Ease of control. The arrangement must lend itself to a reasonably simple and reliable control system with rapid response characteristics.

6. Transient thermal properties. The system should be able to commence effective anti-icing quickly from an initially unenergized state.

7. Structural factors. Depending upon the source of heat, the structure must be adapted to the primary needs of the system. A simple structural and interior layout is desirable.

8. Fundamental means for ice prevention. Other than fluid deicing, four thermal systems are available. Selection of a system must include consideration of the influence of the deicing system on the functional behavior of other components of the airplane.

9. Reliability and safety.

10. Aerodynamic considerations. The velocity distribution over the windshield surface and the amount of water catch will be determined by the air flow characteristics over the nose and upper front fuselage. Thermal requirements are affected directly.

11. Pilot comfort. A high temperature at the inner windshield surface will result in excessive radiant heat on the pilot. Discharging large amounts of warm air in the cockpit enclosure from a hot air system is likewise undesirable.

12. Maintenance and ease of servicing.

13. Fabrication of windshield. A means of ice removal requiring a panel difficult to manufacture and correspondingly costly is to be avoided. Replacement as well as initial cost would be high.

14. Defogging and defrosting. Collection of fog or frost on the interior windshield surface also must be avoided, the simplest means being to maintain the inner surface above the dew point temperature for the given cockpit air humidity, density, and temperature.

Miscellaneous factors arise from other sources. For example, one air line is considering the addition of a thermal deicing system on a modified Douglas C-54 airplane. Since the line figures a revenue loss of over 1,000 dollars for each day the airplane is out of service, a system which can be installed quickly is highly desirable, even though a weight penalty of several pounds is exacted.

A discussion of the possible methods of supplying heat to the windshield will assist in forming a decision as to the most desirable scheme for a given airplane.

Nonelectrical Methods

Two nonelectrical deicing systems are in common use and will be discussed briefly for comparison with electrical methods. The earliest and most widely used method consists of a mechanical wiper which also distributes an alcohol fluid freezing point depressant over the wiping surface. There are definite limits to the deicing ability, and of course no protection at all is afforded after the supply of fluid is exhausted. The system is, however, simple and reasonably light in weight, and may be applied readily to most structural configurations without modification. Curved windshields present an obvious problem. Other advantages and disadvantages are evident and will not be enumerated.

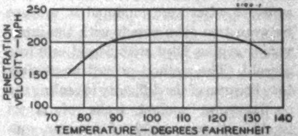

Figure 1. Variation of bird penetration velocity with temperature of a laminated glass and vinyl plastic windshield

Hot air anti-icing has received much attention since 1940.[3] The most commonly used arrangement consists of two panes with an air gap of from 0.1 to 0.2 inch. The heated air is introduced at one edge, passes between the two panels, and is discharged directly into the cockpit, over other windows for defogging, or elsewhere. The inner panel is usually of impact resistant construction and the outer designed for maximum heat transfer consistent with adequate strength. One heated air scheme introduces the air to the windshield on the outside surface in the form of an artificial boundary layer.

The hot air system is reasonably light in weight, although the weight of a second panel is appreciable. Optically, the double windshield is poorer than a single pane arrangement because of the multiplicity of light images, both transmitted and reflected, occasioned by the two additional prime surfaces. A distinct disadvantage is the high temperature and mass of air flow required to provide the necessary thermal energy for combatting a severe icing atmosphere. This may overheat any vinyl plastic present, causing permanent small bubbles to form, and also appreciably reduce the impact strength of a laminated panel. Further, the resulting hot inner surface radiates excessive heat to the pilot's face. The glass may crack because of the high thermal gradients. However, the system has been adopted for numerous military and commercial aircraft, largely because of the relative ease in obtaining heated air, either from waste exhaust heat or primary combustion heaters, compared with the difficulty of procuring sufficient electric power and specially designed windshields.

Embedded Wires

Perhaps the most obvious electrical method of developing heat at the windshield surface is by means of embedded resistance wires. This idea is not new. Rodert describes experiments in which a series of wires were mounted in a space between two glass panels with the intervening region filled with liquid ethylene glycol.[4] The method proved unsatisfactory because of the difficulty in sealing the configuration against leaks of glycol liquid. A windshield of laminar construction with a layer of vinyl plastic sandwiched between two panes of tempered glass conceivably could have wires embedded in either the glass or the vinyl. Because of fabrication difficulties, however, it is impracticable to embed the wires in the glass. One air frame manufacturer has intensively pursued an investigation of panels having the fine wires situated at the interface of the glass and plastic.

The wires must be of a material which can be drawn to a diameter of about one mil or less and still possess adequate tensile strength. The resistance must be correlated with the required deicing power per square foot, wire spacing, and operating voltage. The wire surface should be dull to avoid reflections. Several materials are suitable, such as copper, platinum, copper encased in platinum for strength, beryllium-copper, or nickel. During manufacture the wires are bonded to the vinyl while under tension. Wire buckling and kinking thus is prevented both during panel fabrication and while being heated electrically. The ends are soldered to a bus, usually running horizontally at the top and bottom of the windshield with the wires contained vertically. Visual distraction is less with this arrangement than with the wires mounted horizontally.

A typical design will be described. Beryllium-copper wire one mil in diameter is embedded at or very near the interface between the outer pane and the vinyl. The thicknesses of glass and plastic depend upon the impact strength desired, except that a sufficiently thick front glass pane is required to permit a uniform temperature distribution over the outer face. Three-quarter tempered glass panels 3/16 inch thick, with 1/8 inch to 1/4 inch vinyl are typical dimensions for a windshield 15 inches by 30 inches over-all. Assuming an anti-icing heat demand of 1,740 Btu per hour per square foot with a five-per-cent loss into the pilot enclosure, 535 watts per square foot must be developed by the heated wires. Mounting the wires vertically, and allowing a one-half-inch-wide horizontal bus at top and bottom, the wires will be 14 inches long. One mil beryllium-copper has a resistance of about 60 ohms per foot, and at 27 volts each wire will dissipate 10.4 watts. Thus 60 wires will be necessary per foot of panel, or a wire spacing of 0.2 inch.

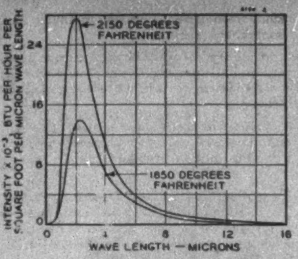

Figure 2. Dependence of radiation intensity upon wave length for a black body

A German windshield, reported by a technical observer of the "Combined Intelligence Objectives Subcommittee" under the British and United States Armies, is described as having 0.015 millimeter (0.6 mil) of copper wire embedded in a plastic middle layer, with a very narrow spacing of the order of 0.05 inch.

The deicing ability of an embedded wire panel should be good, but will be limited by the temperature developed in the vinyl. Exceeding the temperature for maximum impact strength is one consideration; another is a 200-degree-Fahrenheit limit imposed by the tendency of the plasticizer to vaporize above this temperature, forming permanent small bubbles. Values much higher than the critical limit for bubble formation must be developed at the wire surfaces to produce a lower required average vinyl temperature.

The important factors favoring hot wire windshields are the light weight associated with operation on 28 volts, their reliability, and the efficient manner in which the developed heat energy is dissipated.

The disadvantages of the scheme are fundamental and serious. Optically, there are two deterrents. First is the necessity for looking through the closely spaced fine wires with the concomitant eye fatigue. Viewed from a distance of two feet, with eyes focused on objects beyond the windshield, the visual distraction is slight but not insignificant. Second, and of greater importance, is the optical distortion caused by the local heating of the vinyl sheath about each wire. The resulting localized variation of the index of refraction will distort severely objects viewed through the panel. The extent of distortion is proportional to the temperature gradient, and provision of sufficient heat for anti-icing under moderate to extreme conditions may render the windshield optically useless. This condition is tempered by virtue of an outside visibility of zero during icing conditions anyway, but the significant time required for the distortion to fade away after the panel is de-energized may be serious. The distortion produced by the heat necessary in clear air for interior defogging and for impact strength further establishes the seriousness of the situation.

The necessary construction renders the windshield difficult and expensive to manufacture, and the deicing ability is limited by local overheating of the vinyl. If the necessity for supplying heat in clear air conditions (for impact strength and defogging) or the inadvisability of so doing (because of distortion) are disregarded, a complication is introduced by the thermal lag of the system. The time which elapses between the application of power and the inception of ice removal could be serious in an emergency.

Infrared Radiation

Radiant thermal energy directed on a solid body will be absorbed to a varying degree. Most solids are sensibly opaque to infrared radiation, but because of the characteristic similarity between energy in the visible spectrum and in the near infrared spectrum, visually transparent substances such as windshield constituents will pass a certain portion of the energy entering the surface. Because of surface reflection, only a part of the radiation impinging on the surface will enter any material.

The amount of energy reflected or absorbed by a receiver is a function of the wave length of the radiations emitted by the source.[8] The visible spectrum is included within the range of from 0.35 to 0.8 micron. The infrared spectrum is divided arbitrarily into the near infrared spectrum ranging from 0.8 micron to around 25 microns, and the far infrared spectrum ranging above 25 microns. In general, thermal radiation at all wave lengths is emitted by hot solid bodies, the total emission depending upon the temperature, the material, and the condition of the surface. Typical radiation patterns for a body at two temperatures is shown in Figure 2. A small fraction of the total energy is in the visible range but the greatest portion is in the near infrared spectrum. At higher temperatures the peak of the curve is shifted to a lower wave length, the position of the maximum being inversely proportional to the absolute temperature.

The dependence upon wave length of the total energy absorbed by an optically transparent receiver is indicated in Figure 3. Ordinary glass and vinyl plastic readily pass the visible frequencies. The very near infrared spectrum is largely passed by the glass, but strongly absorbed by the vinyl. Thermal radiation at frequencies above 2.8 microns is absorbed by both materials. Special glasses are available which are opaque to visible radiation, but which transmit most of the infrared rays from one to four microns. Conversely, others transmit most of the visible rays, but strongly absorb all wave lengths above 0.8 micron.

An anti-icing scheme using infrared rays is apparent. A grid consisting of electrically heated wires or a series of specially constructed radiant heat lamps are practical as a source. A polished metal reflector will assist in concentrating or distributing the energy on the windshield. An infrared filter may be incorporated, depending upon the amount of visible light developed (a direct function of source temperature) and the necessity for eliminating it. Directing the radiation on the windshield will cause it to rise in temperature, and if sufficient energy is supplied, effective ice prevention may be accomplished.

Most of the heat is developed within the windshield because of the relatively strong absorptive characteristic of the vinyl. The proportional amount for a specific case may be determined from Figure 3. The heat then is transmitted by conduction to the outside. The amount of energy passing on through the assembly may be reduced by using a heat absorbing glass as the outer pane, thus collecting more heat energy in this plane. However, radiation energy passing completely through the windshield will be absorbed by ice or water which may have collected on the surface. Laboratory tests show that a layer of ice so irradiated tends to "mush-up" and melt throughout the layer, a phenomenon readily explained by the penetrating power of the radiation.

A similar layer melted only by heat conducted through the glass displays a tendency to melt away at the glass surface, leaving an ice bridge over the windshield but separated from it by a small air or water filled space.

Suitably locating the source of the infrared rays is one of the major problems presented by this system. Pilot vision must not be impaired, but a reasonably uniform distribution of heat must be secured. These conditions are difficult to meet except for highly sloping windshields, since the incident radiation should strike the windshield at as near 90 degrees as possible. Any solution for windshield configurations similar to the DC-3 or DC-6 is apt to be awkward, and require considerable skill in overcoming the inherent spatial problem. The ratio of source space required to power emanated is of the order of one to two cubic inches per watt. The applicability of radiant heating thus is seen to be greatly dependent upon the structural configuration of the cockpit.

One answer to the space question is to locate the source outside of the airplane. Some form of retractable mounting for hermetically sealed source could be employed in conjunction with a filter to eliminate the visible light. The slope of the windshield may make it desirable to mount the source above the windshield rather than on the nose. This or other factors combined with the high space-to-power ratio again emphasize the fundamental awkwardness of the scheme based on infrared radiation.

Advantages of a radiant heating scheme are light weight and ability to operate on 28 volts. A special windshield design is unnecessary, and the appropriate plastic temperature for satisfactory impact strength may be maintained, although this temperature will not be synonymous with the all anti-icing heat requirements. Indeed, because of the absorbtive capacity of the vinyl in the near infrared spectrum and its low thermal conductivity, an inside source may overheat seriously the inner plane of plastic in an attempt to transmit sufficient heat to the outer glass.

Major disadvantages are the cumbrous installation demands, the difficulty of obtaining even heat distribution over the windshield, and the fire hazard inherent in operating the source at a suitably high temperature. The over-all energy efficiency is poor because of heat losses at the source by convection and conduction plus the loss in a filter should one be necessary. There are also practical limits to the amount of heat energy that can be developed by the source, and to the amount the windshield can absorb without undue temperature gradients within the glass and vinyl sections. The former condition is more of a limitation than the latter, and is related to the space and safety requirements and thermal efficiency. Either factor may make it impossible for the infrared radiation system to provide sufficient protection against the expected icing environment.

Transparent Electric Conductor

A transparent electrically conductive coating on glass, a coating of unique properties, recently has been developed and introduced by a major glass manufacturer.* The importance and application of such a coating for windshield deicing is at once obvious. A coating of this general description is not new; a film of high electrical resistance has been used in the past on glass surfaces for the prevention of electric charge collection, or electrostatic shielding. The new coating, however, exhibits the very important advantages of relatively low resistance and indestructibility. Instead of resistivities of the order of megohms, the new so-called *NESA* coating, can be produced with a resistivity as low as 100 to 120 ohms per square, this being the resistance measured between opposite sides of a square sheet of any linear dimension. Some idea of the thinness of the *NESA* coating is gained by considering that a film of copper having a resistivity of 100 ohms per square would be only seven billionths of an inch thick (assuming that the dimensional resistance properties are maintained for very thin sections, which may not be the case). This is about 0.03 of one per cent of the wave length of light.

The exact nature of the *NESA* film is not known to the author, but its physical properties and limitations can be described. The basic material is applied to a glass surface and processed at relatively high temperatures. It cannot be applied to such materials as Plexiglas (because of the high processing temperature, and possibly other factors). The film has excellent abrasive resistance properties, being of the nature of the glass itself. The coated glass may be partially tempered, but a high temper presents some difficulties. A laminated panel can be made with the coating next to the vinyl plastic if so desired. A *NESA* coating within a laminated panel will introduce a loss in light transmission of around four per cent largely by reflection loss caused by the difference in the index of refraction of the glass or vinyl and the *NESA*. A coating on the outer glass surface will double this loss and occasion more objectionable reflections. The index of refraction of the

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* The Pittsburgh Plate Glass Company.

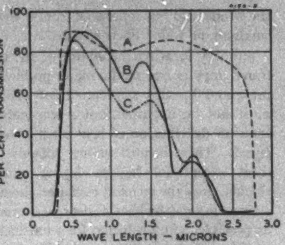
NESA is approximately two. When viewed by transmitted light, the panel imparts no noticeable reflections or color. By reflected light, the reflections from the *NESA* surface are sharp and possess a greenish or reddish caste. The coating appears mottled under close inspection and formed of small splotches reflecting red and green.

Electrical connection is established by means of a narrow opaque bus applied as a coating to the surface of the *NESA* film and usually running lengthwise along opposite edges of a panel. A copper strip may be soldered to the bus and brought out to serve as a terminal connection point.

The uniformity of the *NESA* coating on an aircraft windshield is important, since a nonuniform film will produce localized hot and cold spots. Several large laminated panels, with the coating either on the outer glass surface or next to the vinyl, have been laboratory tested for thermal variance. The results show a maximum deviation from the average of about plus or minus three degrees Fahrenheit for spot checks at points two inches apart over the entire windshield surface. These data were obtained with the horizontally mounted panel in still air with a steady state average surface temperature of approximately 100 degrees Fahrenheit.

A windshield has been designed for a pressurized Douglas airplane which is expected to provide freedom from icing under the most severe anticipated conditions at any operating speed or altitude of the plane. Further, impact resistance of significant excess over the minimum requirement may be incorporated at all times. The panel is approximately 13 inches wide, 28 inches long at the top and 31 inches long at the bottom. A maximum power of 3,100 Btu per hour per square foot is to be dissipated by the *NESA* film which has a resistivity of 120 ohms per square. The heat requirement was computed by methods previously described as that required to anti-ice for the severest expected atmospheric conditions at minus ten degrees Fahrenheit, sea level, and 270 miles per hour, and assuming that the outer surface of the windshield is maintained at 35 degrees Fahrenheit. The average vinyl temperature will be controlled at 110 degrees plus or minus ten degrees Fahrenheit for greatest impact strength, and will not exceed 125 degrees Fahrenheit under this maximum power demand condition.

First considerations in the design were the location of the *NESA* layer, and the glass and vinyl thicknesses. For the most rapid response and most effective deicing, the coating should be put on the windshield outer surface. This, however, introduced the problem of protecting the bus from abrasion, a safety hazard because of the exposed coating, and necessitated other means, such as another interior coating, for maintaining a warm vinyl and for defogging. The best compromise was a *NESA* coating on the inner surface of the outer pane. The power of 3,100 Btu per hour per square foot will be conducted through the outer pane at a negative temperature gradient of 465 degrees per inch, neglecting edge conditions. A glass panel 3/16 inch thick will transmit this thermal energy at a temperature difference of 87 degrees Fahrenheit, thus requiring 122 degrees Fahrenheit at the vinyl-glass interface, a figure which is approximately the upper limit for maximum impact strength. A thinner front pane would be satisfactory thermally and more desirable from the standpoint of rapid transient response, but strength and glass temper considerations lead to the selection of the greater thickness. The same thickness was adopted for the inner pane, and 1/8 to 3/16 inch adopted for the vinyl. The temperature difference across the vinyl under operating conditions will be only a few degrees.



Pittsburgh Plate Glass Company

Figure 3. Transmission characteristics as a function of wave length

A—Polished plate glass, 6.17 millimeters
B—Vinyl plastic, 3.1 millimeters
C—Laminated panel, 0.120-inch vinyl plastic encased between two $^{7}/_{64}$-inch glass panes

The described windshield incorporates a 5/16-inch bus running horizontally at the top and bottom edges behind the mounting frame. The heated area is 2.65 square feet requiring 2.4 kw at 3,100 Btu per hour per square foot. This amount of power will be developed by 350 volts for a coating resistivity of 120 ohms per square.

Control may be effected by maintaining a vinyl temperature appropriate for high impact resistance. At 110 degrees Fahrenheit nearly all icing environment can be countered; for the most severe conditions the control setting can be increased to 120 or 125 degrees Fahrenheit. Only a minor sacrifice in impact strength will be introduced by maintaining the temperature in the upper range, and full anti-icing ability will be provided at all times. Or, if sufficient strength is developed at a lower temperature, say 70 degrees Fahrenheit, control could be maintained at this point with a corresponding improvement in pilot comfort in a warm cockpit. Several schemes for control have been advanced which will not be enumerated here. The design will depend largely upon the following three factors:

1. The source of electric power and type of generating equipment.
2. The type of temperature sensing element incorporated with the windshield.
3. The transient thermal response of the windshield and associated components.

At the maximum flying speed in cold clear air, sufficient power must be supplied to the windshield to maintain the optimum vinyl temperature. The previously quoted figure of 3,100 Btu per hour per square foot will suffice to maintain a vinyl temperature of 120 degrees Fahrenheit for an outside air temperature of minus 70 degrees Fahrenheit at 270 miles per hour.

Provision for stand-by in case of control or generating equipment failure may be obtained by using two control units and two alternators or inverters. Alternating current is the most logical type of power because of the high voltage requirement and absence of disturbing unidirectional fields. One generator may be of a capacity that is sufficient to do the whole job, with a second machine for stand-by; or two generators of half capacity may be used with a switching arrangement to allow application of power to the pilot's windshield system only, should his generator or control fail. Thus at least half of the pilot's enclosure could be kept ice-free in an emergency, and for light icing conditions the one generator could deice the entire windshield system.

The transparent electrically conducting coating of low resistance appears to be an ideal solution to the windshield icing and impact proofing problem. The maximum amount of heat required for the worst conditions may be supplied easily and used efficiently. The system lends itself readily to control both from anti-icing and impact resistance considerations. A *NESA* coated windshield possesses satisfactory optical characteristics at all times. It is relatively simple to manufacture, not subject to failure in operation, and alleged

Moisture Equilibrium Between Gas Space and Fibrous Materials in Enclosed Electric Equipment

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THE USE of an inert gas in enclosed electric equipment has increased considerably in recent years. In transformers an inert gas is used in the space provided above the transformer fluid to allow for its volumetric expansion with temperature rise. There the function of the gas is to provide an inert medium by which the pressure at the surface of the transformer fluid is kept slightly above atmospheric. Thus the fluid is not exposed to the moisture and oxygen of the air, as it is with the so-called "open-breather" or "oil-conservator" types unless the latter communicate with the outside atmosphere through chemical agents. In gas pressure cables, however, the inert gas is used at pressures up to approximately 15 atmospheres in order that the cables may operate successfully at electric stresses that are higher than otherwise would be possible.[1–3]

The purpose of this paper is to call attention to a phenomenon that occurs in enclosed electric equipment in which both an inert gas and fibrous material are present. The systems involved tend to establish an equilibrium between the moisture content of the fibrous materials and that in the gas space. Consideration, therefore, should be given to the moisture content of the fibrous materials present as well as that of the gas itself. A quantitative method is suggested for predetermining the equilibrium condition for both unimpregnated and impregnated fibers. It is believed that use of the method may assist in the solution of manufacturing and operating problems.

Field Experience and Data

The increasing of the moisture content of dry nitrogen by the transfer of moisture from fibrous materials has been observed in The Detroit Edison Company in transformers and in the 120,000-volt gas-pressure cable.[1–3] This increasing of the moisture content of the gas can be of practical significance only if the dew point of the gas is reached at local cool spots over considerable periods of time, thus causing progressive transfer of moisture from the warmer fibrous materials, and if the resulting condensate drops upon a vulnerable part of the equipment. Any condensation that takes place as a result of an over-all cooling of the equipment or of a cool spot of short duration is of no significance because the water content of the gas is relatively low.

During severely cold weather, it is probable that some condensation occurs in all of the transformers that have been investigated. Moreover, it is probable that, in certain transformers, condensation occurs during the cool part of the day nearly every day of the year. Apparently the condensate usually drops in locations where it does no harm. Condensate has been found, however, on the terminal board inside transformers. Whether or not service failures actually have been caused by condensate on vulnerable parts is unknown, but there seems little question that such condensation constitutes a potential source of trouble.

In the case of the 120,000-volt gas-pressure cable, in which the paper-insulated unsheathed cables rest on the bottom of a steel containing-pipe, nearly all locations are vulnerable to attack by liquid water should condensation occur at local cool spots. Such cool spots might occur at a terminal, a length that crosses

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to be free from deterioration caused by aging, sunlight, overheating, and abrasion. During operation the inner surface of the windshield is not excessively hot, which improves pilot comfort by minimizing the radiant heat. The installation should prove reliable and safe. Structurally, a clean design should result, the only complication being the means for attaching the power wires and control elements.

There are disadvantages to the scheme, however. Most important is the weight, which may be high because of the necessity for using high voltage alternating current delivered by inverters or engine driven alternators. The control system adds significant weight, although a unit to effect the same order of regulation in other ice prevention schemes would be of comparable weight. Inability to operate the *NESA* film directly from the ship's 28-volt power supply and provide the necessary power is a disadvantage. The requisite high voltage introduces a safety hazard, although placing the *NESA* coating next to the vinyl will tend to mitigate the danger. Windows or shields made of plastics such as Plexiglas cannot be deiced or defogged by this method since the *NESA* coating cannot be applied to their surface, at least not by present techniques.

Conclusions

1. Aside from anti-icing, an additional prime reason for heating aircraft windshields is evident. For a given weight and construction, vastly improved impact resistance can be obtained by using a laminated panel with the vinyl plastic maintained at about 110 degrees Fahrenheit.
2. The method of heating a windshield will depend greatly upon the structural configuration of the cockpit and of the windshield system. Availability of heated air and (or) electric power are equally important factors.
3. Use of the transparent electrically conductive coating on glass is the most desirable electrical method for providing heat at the windshield. Indeed, unless operating experience proves otherwise, this method appears to be suited ideally for the purpose of windshield anti-icing. For most airplanes, it ultimately should supplant other presently known systems.

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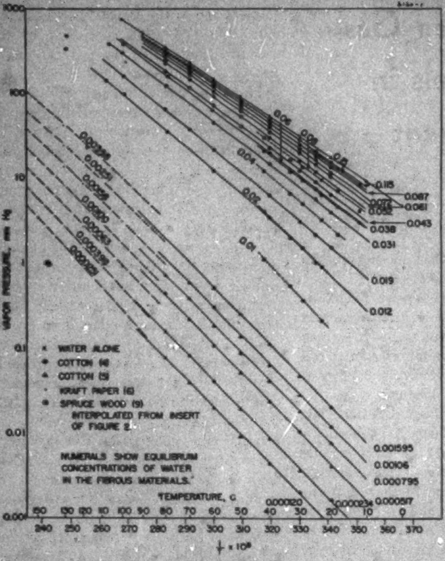

Figure 1. Vapor pressure of water alone and in fibrous materials; data from literature

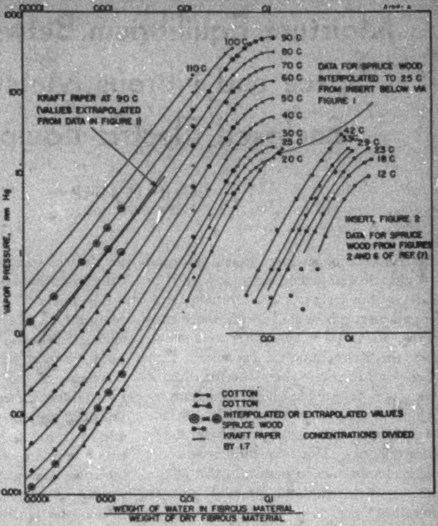

Figure 2. Equilibrium between water concentrations in fibrous materials and vapor pressure at selected temperatures

a canal, or a location in proximity to a water main. For this reason when it was found during the progress of the installation that the moisture content of the dry nitrogen gas, which was introduced immediately after the cable-pulling operation, was beginning to increase, steps were taken co-operatively by the manufacturer and user to determine the cause and corrective measures necessary. These efforts were completely successful as demonstrated both by the data herein contained and by the performance of this circuit since it was energized in December 1941. The data concerning this cable are presented in this paper only because they indicate that a predictable equilibrium exists between the moisture content of the fibrous material, even when impregnated, and that of the gas space.

As has been described elsewhere[1,3] the 120,000-volt gas-pressure cable is of preimpregnated-tape construction; that is, the cable was taped in an air-conditioned room with tapes that previously had been impregnated and from which the excess impregnant had been removed. Over the taped insulation two layers of metallized-paper shielding tapes were applied and over these a thin bronze tape intercalated with an impregnated muslin tape. This was followed by two layers of paraffin-impregnated canvas* to provide mechanical protection during installation. After a spiraled skid wire was wrapped over the canvas and the assembly coated with petrolatum for lubricant during installation, the cable was enclosed in a temporary lead sheath that was stripped within a few feet of the entrance tube to the pipe line as the cable was being installed.

When the dry nitrogen, which was introduced into the steel pipe line with the cable and held at slightly over atmospheric pressure by temporary plugs, began to show the presence of moisture, the source of the moisture was sought. The pipe was known to be dry. No water was found in the petrolatum grease but from 2.4 to 5.0 per cent of water** was found in the impregnated muslin and canvas tapes,* based upon the weight of dry fabric. The average for 14 samples was 3.3 per cent. Two steps were taken to minimize the moisture content of the muslin and canvas tapes.

1. The manufacturer modified his process in such a manner that the moisture content of the muslin and canvas tapes on further shipments of cable was reduced to an average of 1.67 per cent as judged by tests on six samples in which the content varied from 1.0 to 2.0 per cent.

2. Evacuation of the completed line, which is a normal procedure in the installation of this type of cable, was extended for eight days while the cable was heated by conductor heating to give a tape temperature of between 40 and 42 degrees centigrade.[8] The water removed was collected in a carbon-dioxide-alcohol trap and measured. From the average water concentrations and the weight of water removed, the average moisture content of the protective tapes as the line was placed in operation was calculated to be 1.73 per cent. Presumably, most of the water that was removed came from the wetter tapes.

In order to determine whether these efforts to prevent the possibility of condensation were successful, periodic determinations were made of the moisture concentrations† or the dew points of the gas space and of the temperature at various points along the line. During the first year of operation the dew points rose until late summer when at the more moist locations, which contained cable not given the special factory treatment, they were above the winter operating temperature of the pipe line. In winter, however, the dew points fell to values safely below the winter operating temperature.

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* This type of construction no longer is used by the manufacturer and hence it does not follow that dew point problems are inherent in other gas-pressure-cable installations.

** As determined by Dean-Stark apparatus and boiling toluene.[4]

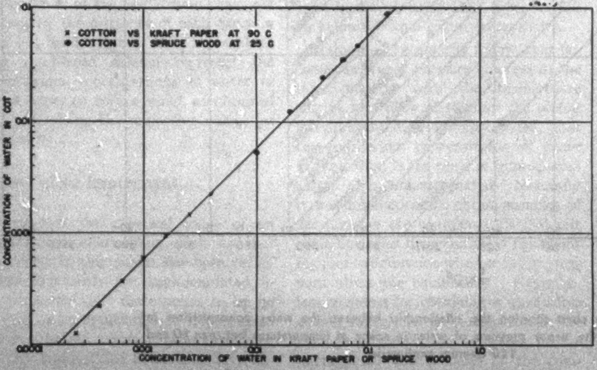
† Originally the moisture concentrations were determined gravimetrically at atmospheric pressure and the results calculated to line pressure. Later dew points were determined at line pressure using a dew-point apparatus.

Inasmuch as the rise and fall, with temperature change, of the vapor pressures that corresponded to the several dew points or moisture concentrations followed the vapor pressure rule for liquids, within experimental error, it became evident that an equilibrium existed between the moisture content of the gas space and that of the tapes. This behavior, coupled with the change in dew points of transformers with temperature change, led to a search of the literature to determine whether quantitative data existed concerning this equilibrium for the fibrous materials generally used in the electrical industry. Literature on the subject revealed that although there are many articles describing the water concentrations of various fibrous materials under various relative humidities, only a few show the effect of temperature change. Fortunately two such sets of data were available for cotton and one each for kraft paper and spruce wood. From these data an equilibrium chart has been prepared that is useful for predicting and interpreting the behavior of the moisture within electric equipment containing fibrous materials and a gas space.

Preparation of the Equilibrium Chart

Figure 1 shows the data as obtained from the literature. In this figure the vapor pressures, on a logarithmic scale, are plotted against the reciprocals of the absolute temperatures on a linear scale. The top curve is for water alone. The long solid curves below the top curve represent the data of Urquhart and Williams[5] for cotton containing over one per cent of water. The data for each concentration are represented as a continuous curve instead of as the two intersecting straight lines used by the original authors. As thus represented, the curves for the higher concentrations are nearly parallel to, and have nearly the same curvature as, the curve for water. With diminishing concentration of water, the curves become more divergent and their slopes approach that of the curve for ice. This curve for ice is not shown in Figure 1. The solid curves at the lower right hand corner of the figure represent the data of Neale and Stringfellow[6] for cotton of low moisture content. These curves are straight, parallel lines having the slopes of the vapor pressure curve for ice. The curves that appear to extend as broken lines from the curves representing Neale and Stringfellow's data represent the data of Houtz and McLean[7] for kraft paper of low moisture content at elevated temperatures. The data for the samples having the higher moisture concentrations are best represented by curved lines whereas the data for the samples of lower concentrations fall along straight lines as shown. The latter are not parallel with each other. For the data of Urquhart and Williams, the slopes of the curves become steeper as the moisture concentrations of the samples represented diminish. The fourth set of data is that of Pidgeon and Maass[8] for spruce wood. These data appear in the same region as those of Urquhart and Williams. Because the temperature range for the investigation on spruce wood was narrow, it is not clear whether the data are best represented by straight or curved lines. The slopes of the lines are approximately the same, however, as those of the adjacent curves for cotton.

Figure 3. Comparison between the concentration of water in cotton and that in kraft paper or spruce wood required to give the same vapor pressure at the same temperature



In using the four sets of data to prepare a general equilibrium chart, no effort was made to distinguish between absorption and desorption data. The difference, where shown by the original authors, is small as compared with the effect of temperature change, which is the primary concern of this paper, and probably is also small as compared with the effect of the previous history of the samples.[9]

The data for the equilibrium chart for cotton were obtained essentially by interpolation, for selected temperatures, of the vapor pressures that correspond to chosen concentrations of moisture. The principal object of the interpolation, of course, was to bridge the gap between the data of Urquhart and Williams and those of Neale and Stringfellow in the concentration range between 0.0016 and 0.012 gram of water per gram of dry cotton. Interpolation was carried out by means of Figure 2, where the vapor pressures and concentrations of water in cotton are both plotted on logarithmic scales for selected temperatures to give the long S-shaped curves that represent both the data of Urquhart and Williams and those of Neale and Stringfellow. In order to obtain the 90-degrees-centigrade curve of Figure 2, the data of Neale and Stringfellow were extrapolated from 80 to 90 degrees centigrade by extending the straight lines as shown in Figure 1.

Because only one set of data was available for kraft paper‡ and that at high temperatures only, interpolation similar to that described for cotton could not be carried out. Instead, the curves in Figure 1 for the data of Houtz and McLean were extrapolated downward from 100 to 90 degrees centigrade, and the resulting values of vapor pressure were plotted against their respective concentrations in Figure 2. The straight line that resulted when the extrapolated values for kraft paper at 90 degrees centigrade were plotted on the double-log scale of Figure 2 lies slightly below the curve for cotton at 80 degrees centigrade as shown. Next, it was reasoned that inasmuch as kraft paper often is made from spruce pulp, possibly the equilibrium in kraft paper and spruce pulp might be related. For this reason the data of Pidgeon and Maass,[8] shown in the insert of Figure 2, were in-

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‡ Vincent and Simons[10] evidently investigated the field over a wide range of temperatures and moisture concentrations but unfortunately did not include in their publication data suitable for preparing an equilibrium chart.

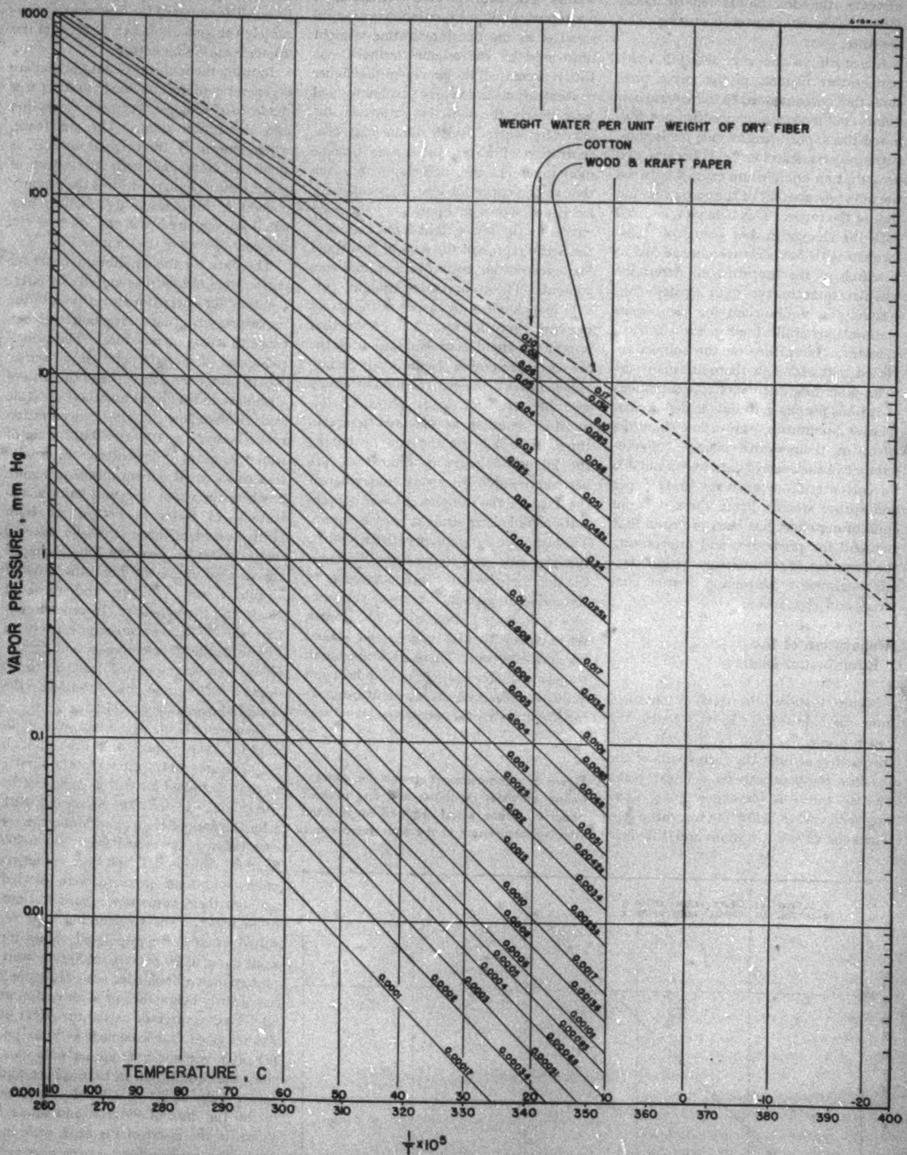

Figure 4. Equilibrium chart showing the relationship between the water concentration in fibrous materials and the vapor pressure of water in space at temperatures between 10 and 110 degrees centigrade

terpolated by the use of Figure 1 to 25 degrees centigrade. Finally, the concentrations of water in cotton required to give the same vapor pressure at 90 degrees centigrade as are given by selected concentrations of water in kraft paper and those concentrations required to give the same vapor pressure at 25 degrees centigrade as are given by selected concentrations of water in spruce wood were plotted on a double-log scale as shown in Figure 3. The straight line through these points indicates that the relationship between the water concentrations that are required to produce the same vapor pressures in cotton and in either kraft paper or spruce wood is simply the ratio 1 to 1.7.

By dividing the concentrations of water in kraft paper, as given in Houtz and McLean's data, by 1.7, an estimate was made of the vapor pressures that would result from low concentrations of water in cotton at 100 and 110 degrees centigrade. These data were combined with the experimental data of Urquhart and Williams in that temperature range to produce the top two curves of Figure 2.

The equilibrium chart was prepared by plotting the vapor pressures for cotton, as given from the curves of Figure 2 for selected concentrations, on a logarithmic scale against the reciprocals of the absolute temperatures, similarly to Figure 1. This chart is shown in Figure 4. All of the curves are extrapolated from 20 to 10 degrees centigrade. Those for concentrations above 0.03 gram of water per gram of dry cotton also are extrapolated for temperatures above 80 degrees centigrade. For this extrapolation, the gentle curvature of the vapor pressure curve for water, shown as the broken line constituting the top curve, was followed.

The equilibrium concentrations of water in cotton that were selected for the preparation of the equilibrium chart are shown by the numerals of each curve in the left-hand column. The numerals in the right-hand column represent the equilibrium concentrations of water in kraft paper or spruce wood, as obtained by multiplying the respective values for cotton by 1.7.

Effect of an Impregnant

Although the practical value of an equilibrium relationship, such as that described in this paper, has been recognized previously for unimpregnated fibrous materials,[10] there seems to be no record of its application to impregnated fibers. Indeed, to many who have tried to dry impregnated fibers that have absorbed moisture, it might appear that impregnated fibers behave entirely differently from unimpregnated ones. The data for the paraffin-impregnated cotton protective tapes of the 120,000-volt cable previously described provide a comparison between the behavior of commercially made American impregnated-cotton tapes and that of the two sets of laboratory-prepared unimpregnated-cotton fibers made in England.

The comparison between the impregnated and the unimpregnated cotton samples is given in Figure 5. The solid lines in this figure, and the numerals arranged in a column upon them, are reproduced for cotton from Figure 4. The crosses represent the average conditions at three locations at which the higher moisture concentrations were found in the gas space as the average temperature of the pipe line rose and fell with seasonal and load change. The low-temperature point was determined while the line was not in operation. The circles represent the average conditions at two locations at which the lower moisture concentrations were found. It is evident that, regardless of whether the fibers are impregnated or not, the change of the vapor pressure of water in the gas space is essentially the same function of temperature change. It is also evident that although equilibrium had not been reached when some of the data on the cable line were obtained, especially those taken in September 1942, the moisture concentrations in the cotton, as indicated by the chart for different dates and temperatures, are nearly the same. Even including the September 1942 data, the indicated concentrations ranged only between 2.2 and 2.6 per cent for the tapes in the more moist regions and between 1.2 and 1.5 per cent for those in the drier locations. Data taken during the initial operation of the cable in the winter of 1942, however, fall far below the equilibrium values shown.

It does not necessarily follow, that because the vapor pressure changes in the same manner with the temperature change regardless of whether the cotton is impregnated or unimpregnated, that the equilibrium concentration of water in the cotton is the same in impregnated cotton as in unimpregnated. Naturally it was impractical to obtain samples of the tapes in the completed 120,000-volt cable line after the evacuation treatment in order to determine whether the impregnant alters the equilibrium. Nevertheless, a means for obtaining a qualitative answer exists. As previously stated, the average concentration of water in the impregnated cotton tapes of the completed line, as calculated from the analytical results and from the quantity of water removed, was 1.73 per cent. The points indicated by dots in Figure 5 show the concentrations of water in cotton at each of the 30 joints at which the samples were taken. These were determined with the equilibrium chart from the average temperature of the line and the vapor pressures that correspond to the dew points that were determined in September 1944. The values, which because they were determined by the chart are for unimpregnated cotton, ranged between 0.9 and 2.65 per cent and averaged 1.68 per cent. The agreement between 1.73 and 1.68 per cent is much closer than the analytical methods warrant. The agreement indicates that it is safe to use the equilibrium chart for impregnated cotton at least in the solution of practical problems in which an error of perhaps plus or minus two in the second significant figure can be tolerated.

No data are available to show whether impregnation with insulating oil changes the equilibrium temperature for kraft paper. The data for the paraffin-impregnated cotton support the belief that although the impregnant greatly slows down the diffusion process by which equilibrium is attained, the equilibrium concentration of water in a fibrous material is not affected greatly by an organic impregnant. It seems probable that the equilibrium conditions shown for unimpregnated kraft paper in Figure 4 will apply approximately for oil-impregnated kraft paper as well.

Use of the Equilibrium Chart

The following examples are intended to illustrate a few of the uses that may be made of the equilibrium chart in the electrical industry. The first three of these examples concern cable of the type discussed.

EXAMPLE 1

Did the precautionary treatments of protective tapes (previously described) effect any substantial reduction in the dew point of the gas?

This question is answered by the information in Figure 6, which also shows the method of using the equilibrium chart. As previously stated, before the drying treatments the maximum moisture concentration was 5.0 per cent, and the average 3.3 per cent. After the factory treatment on the part of the cables and evacuation of the entire line, the maximum concentration as determined from the dew points was 2.65 per cent and the average concentration from both the dew point measurements and the calculated residual water concentration was

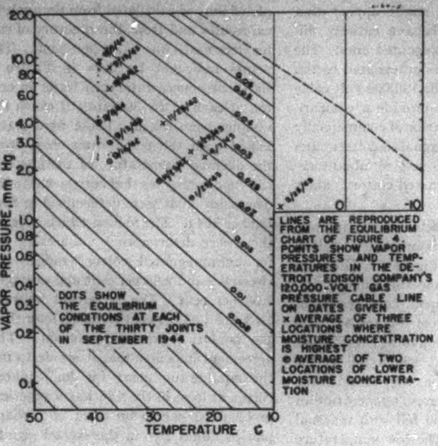

Figure 5. Similarity in equilibrium behavior between impregnated and unimpregnated cotton

1.7 per cent. The dew point of the gas space corresponding to a point that represents a given moisture concentration and selected temperature of the fibrous material is obtained by drawing a horizontal line from that point to the curve for water as shown by the arrows in Figure 6, and obtaining the dew point temperature as given by the point of intersection. Thus the chart shows that the dew point in the vicinity of tapes having the maximum moisture content of 5.0 per cent would be 26 degrees centigrade when the summer temperature of the cable line averages 38 degrees centigrade, and nine degrees centigrade when the winter temperature of the line averages 20 degrees centigrade. Thus it is shown that the drying treatments reduced both the maximum and the average dew points by approximately 15 degrees centigrade over the entire operating temperature range of the line, as shown.

EXAMPLE 2

What concentration of water would a cotton protective tape acquire if the cable on which the tape is wrapped stands, until equilibrium takes place, in a room air conditioned to a relative humidity of 20 per cent at 30 degrees centigrade, and the cable thereafter is sheathed?

A 20-per-cent relative humidity at 30 degrees centigrade is equivalent to a moisture content of 6.0 grams per cubic meter which corresponds to a vapor pressure of 5.7[*] millimeters of mercury. A point representing 5.7 millimeters and 30 degrees centigrade on the chart indicates a moisture content of 2.6 per cent based upon the weight of dry unimpregnated cotton.

EXAMPLE 3

What concentration of water would preimpregnated kraft paper tapes acquire if they were exposed in the air-conditioned room described in example 2 long enough for equilibrium to be reached?

The point representing 507 millimeters of mercury and 30 degrees centigrade on the chart indicates a moisture concentration of 4.4 per cent in kraft paper.

The foregoing example does not imply that the moisture content of a preimpregnated paper insulation is high just because the equilibrium moisture content is high for the humidities that feasibly may be maintained in the taping room. Under conditions of normal operation the time of exposure is too short to allow significant absorption. For example, the average moisture content of the insulation of the cable previously mentioned was not 4.4 per cent but only 0.055 per cent, which is essentially the same as that for other well-made cables impregnated in a conventional manner. An equilibrium value that low would require, according to the chart, the relative humidity at 30 degrees centigrade to be of the order of 0.017 per cent. Obviously, the low rate of moisture absorption is of much more importance to the manufacturer of this type of cable than is the equilibrium concentration.

The two examples that follow illustrate the use of the equilibrium chart for problems involving condensation in transformers. For given equilibrium moisture content in the fibrous material, most of which is of wood origin, the probability of condensation is greater than for gas-pressure cable. The reason is that the top of the transformer casing may be only a few degrees above outdoor temperature while the fibrous material is at the oil temperature, which may reach a daily maximum of 50 to 60 degrees centigrade. The equilibrium for transformers is more complicated than for the paraffin-impregnated tapes because the amount of water in the large mass of oil is not negligible as compared with that in the much smaller mass of fibrous materials. It is probable that a definite relationship also can be established for the equilibrium between the moisture content of oil and that of the gas space. Until such information is available, however, it is not possible to estimate the water content of the oil from dew point data. In the following the oil is considered only as a barrier that markedly reduces the rate at which equilibrium becomes established between the moisture contents of the gas and fiber.

EXAMPLE 4

The gas in a new transformer had a dew point of ten degrees centigrade when operating at an oil temperature of 50 degrees centigrade while the outdoor temperature was well above the dew point of the gas. What was the probable moisture content of the kraft paper insulation within it?

A dew point of ten degrees centigrade corresponds to a vapor pressure of 9.2 millimeters of mercury, which at 50 degrees centigrade is shown by the chart to be in equilibrium with a moisture content of approximately 2.7 per cent in the insulation.

EXAMPLE 5

The gas in a reconverted oil-conservator type of transformer had a dew point of 20 degrees centigrade when operating at an oil temperature of 42 degrees centigrade while the outdoor temperature was well above the dew point of the gas. How much water would have to be removed from the fibrous insulation of the transformer in order for the gas to have a dew point of minus ten degrees centigrade while the oil temperature was 50 degrees centigrade?

A dew point of 20 degrees centigrade, or a vapor pressure of 17.5 millimeters of mercury, indicates an equilibrium moisture concentration of 5.4 per cent at 42 degrees centigrade, according to the chart. A dew point of minus ten degrees centigrade or 1.95 millimeters of mercury indicates a concentration of 1.05 at 50 degrees centigrade. Thus it would be necessary to reduce the water content by approximately 4.35 per cent based upon the weight of the paper fibers. For a transformer containing 3,000 pounds of such insulation, it therefore

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[*] Because the concentration of water in grams per cubic meter and the vapor pressure are nearly the same numerically up to 50 degrees centigrade, only a small error is introduced by multiplying the vapor pressure shown on the chart by the relative humidities to obtain the equilibrium vapor pressures. In this case, instead of 5.7 millimeters, a value of 20 per cent of 32 or 6.4 millimeters would be obtained.

would be necessary to remove approximately 130 pounds or 15.6 gallons of water which does not include the amount that would have to be removed from the oil to reach the same equilibrium conditions.

Vapor Barriers

Consideration of the information given by the equilibrium chart leads to the conclusion that, if no vapor barrier existed between the cotton and the kraft paper tapes of the cable previously mentioned, equilibrium ultimately should become established between the moisture content of the cotton and kraft tapes with resultant reduction in the dew point of the gas space. The two kinds of tapes are present in the fiber-weight ratios of 22 to 485. Considering the region in which the final concentration of moisture in the cotton tapes was highest (2.65 per cent), if equilibrium were established with the kraft types, which had an average concentration of 0.055 per cent, the final moisture concentration of the cotton tapes would be 0.1 per cent and of the kraft tapes 0.17 per cent. The equilibrium vapor pressure at a temperature of 38 degrees centigrade then would be, according to the chart, approximately 0.05 millimeter of mercury, and the corresponding dew point approximately minus 45 degrees centigrade. Although establishment of the equilibrium between the moisture contents of the cotton and kraft tapes would be expected to be slow, there was no evidence that any exchange had taken place in over $2\frac{1}{2}$ years of operation. This is remarkable when it is realized that the conditions are the same as though the cable had been exposed continuously for over two years to a relative humidity of approximately 30 per cent. Furthermore, exposure for several months of unsheathed samples of the cable to Detroit's humid summer atmosphere out of direct contact with rain resulted in no apparent increase in moisture content of the kraft insulation. Of three samples of cable so exposed before the line was completed, the moisture concentrations were 0.01, 0.01, and 0.05 per cent, respectively, all of which values happened to be less than the average for unexposed samples. The author's interpretation of the reason the insulation does not become moist is that the metallized-paper shielding tapes bar the diffusion of water vapor into the insulation. This fact is believed to be of further practical significance, for no longer can it be said that plastic cable sheaths cannot be used for oil-impregnated cables on the basis that all such sheaths, although they provide a barrier for liquid water, are too permeable to water vapor. It seems entirely possible that if an elastomer can be developed to have suitable life in contact with oil on one side and ground water and air on the other, it should be a suitable sheathing for cables provided the outside of the insulation of each conductor is shielded by metallized paper or thin metal tapes. The latter without some sort of sheath are not adequate because they will not keep out liquid water nor provide mechanical protection. Use of a corrosion and abrasion-resistant elastic material that can be applied competitively with lead seems worthy of investigation, particularly for solid type cables. The elaborate "sandwich" scheme of Rihl and Heering[11] to use a plastic sheath seems unnecessary.

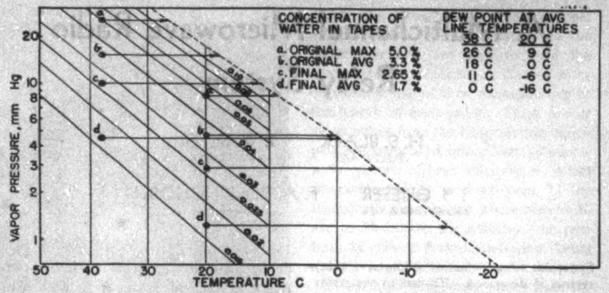

Figure 6. Use of the equilibrium chart in estimating the reduction in dew point that was effected by reducing the water concentration in impregnated cotton tapes

Conclusions

An equilibrium chart has been prepared from data from the literature to relate the equilibrium concentration of water in cotton or kraft paper with the vapor pressure of water, over a temperature range of 10 to 110 degrees centigrade. The concentrations of moisture covered by the chart range from 0.01 to 10 per cent for cotton and 1.7 times those values for kraft paper.

Although the equilibrium chart was prepared from data for unimpregnated fibers, data taken on The Detroit Edison Company's 120,000-volt gas-pressure cable line show that the chart is also valid for impregnated materials, at least for cotton impregnated with paraffin.

Examples are given of the use of the chart in predicting the equilibrium moisture content that fibrous materials will acquire when exposed to atmospheres of various relative humidities and temperatures, in predicting the dew points of the gas in enclosed electric apparatus, and in employing dew point measurements as a nondestructive analytical tool for determining the moisture content of fibrous materials in electric equipment.

Metallized-paper electrical shielding tapes placed over fibrous materials are shown to be remarkably effective in preventing diffusion of water into these materials. The significance of this phenomenon in connection with the use of an elastomer as a sheathing material, particularly for solid-type cables, is pointed out.

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A Multichannel Microwave Radio Relay System

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Synopsis: An 8-channel microwave relay system is described. Known to the Army and Navy as *AN/TRC-6*, the system uses radio frequencies approaching 5,000 megacycles. At these frequencies, there is a complete absence of static and most man-made interference. The waves are concentrated into a sharp beam and do not travel along the earth much beyond seeing distances. Other systems using the same frequencies can be operated in the near vicinity. The transmitter power is only one four-millionth as great as would be required with nondirectional antennas. The distance between sets is limited but by using intermediate repeaters communications are extended readily to longer distances. Short pulses of microwave power carry the intelligence of the eight messages utilizing *pulse position modulation* to modulate the pulses and *time division* to multiplex the channels. The eight message circuits which each *AN/TRC-6* system provides are high-grade telephone circuits and can be used for signaling, dialing, facsimile, picture transmission, or multichannel voice frequency telegraph. Two-way voice transmission over radio links totaling 1,600 miles, and one-way over 3,200 miles have been accomplished successfully in demonstrations.

The *AN/TRC-6* is a combined transmitter and receiver designed specifically for radio relay purposes and includes multiplex facilities for providing eight two-way high-grade message circuits between points separated by an unobstructed optical path. Acknowledgment is made of the impetus and assistance given in this development by those concerned in a somewhat similar British development (British Wireless Set Numbet 10) conducted by the Ministry of Supply on behalf of the British Army.

In military use, one *AN/TRC-6* set usually is transported by truck, and can be set up and placed in operation on a favorable site in a few hours. Figure 1 depicts an arrangement used by the Army in tactical movements, with the operational units housed in a truck. All of the microwave equipment is atop the 50-foot tower. Simultaneous two-way communication is obtained by the use of different radio frequencies for the two directions of transmission. Using separate antenna systems, the microwave transmitter and receiver-converter are mounted directly behind their respective 5-foot parabolic reflectors.

A sharply focused and highly directive microwave beam whose frequency is nearly 5,000 megacycles is used to carry the intelligence of the eight messages. This extremely high frequency affords communication channels free from static and most man-made interference. By the use of beamed radiation and reception, the transmitter power is only one four-millionth as great as would be required with nondirective antennas. Sharply beamed transmission combined with the absence of external noise means a small amount of transmitter power is adequate for communication over optical paths of considerable length. A peak power of a few watts serves for jumps as great as 100 miles. This power is obtained from a small tube no larger than an ordinary radio receiving tube, operating as a reflex velocity modulated oscillator ("reflex klystron").

Because of the sharp beam, because of the type of modulation used, and because transmission must be over an unobstructed optical path other sets using the same frequency can be operated in the near vicinity.

The sets were designed to be used in pairs to form a radio repeater, and it is practical to operate a considerable number of radio links in tandem. Because of the line-of-sight over which the system operates, the distance between sets is limited by the curvature of the earth, but by using intermediate sets as repeaters, communications are extended readily to hundreds and even thousands of miles.

When the set is used for military purposes, essential units are supplied in duplicate, as indicated by Figure 2 which is a close-up of the operational units. The portable test oscilloscope is for conveni-

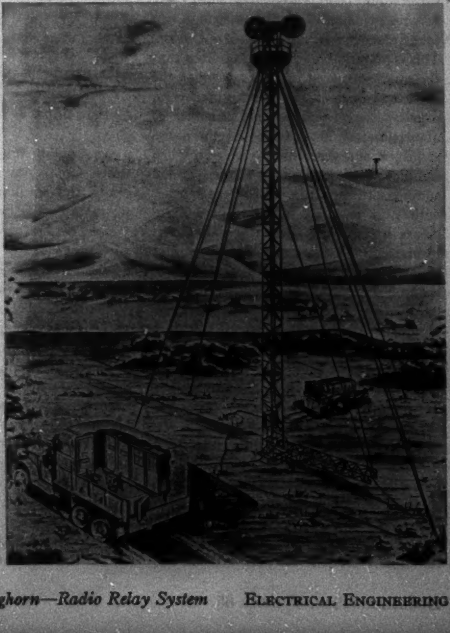

Figure 1. *AN/TRC-6* with operational units in a truck

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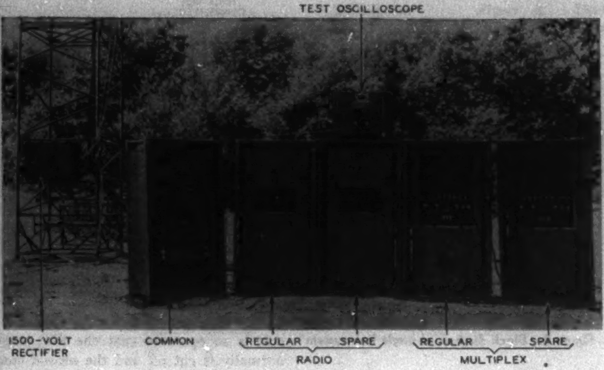

Figure 2. *AN/TRC-6* set up as a field terminal

ence in monitoring. A high voltage rectifier for the transmitter is located at the bottom of the tower.

An optional and improved arrangement representing a recent development and known as *AN/TRC*-6 (*XC*-4) is now available. This improved equipment features one antenna instead of two, a lighter weight tower presenting a less conspicuous target, and the location of the microwave transmitter and receiver-converter on the ground instead of atop the tower. The radio transmitter and receiver-converter connect through wave guide filters to a single antenna system also located on the ground. The parabolic ground reflector (Figure 3) is beamed on a wire mesh 48-degree plane reflector atop the lightweight 50-foot mast. The extra loss introduced by the single antenna system is about three decibels.

Unlike conventional communication equipment which transmits a continuously modulated wave, *AN/TRC*-6 makes use of pulse modulation. In such a system, the transmitter is caused to emit short pulses of microwave power. These microwave pulses are substantially constant in amplitude and frequency. Eight 1-microsecond pulses (Figure 4), one for each channel, are transmitted in sequence and are preceded by a 4-microsecond synchronizing pulse called a marker. This sequence of marker plus eight channel pulses repeats itself 8,000 times a second and is called a frame.

The intelligence of each channel is conveyed by varying the exact position in time of the channel pulses. The phrase pulse position modulation has been applied to this method. The eight channels share the operating time of the transmitter, each using the radio path in turn. Such a sharing process utilizes a multiplexing principle that for many years has been known as time division. Pulse modulation, therefore, not only provides for position modulating the pulses but also permits multiplexing the channels by time division. No detectable distortion of the recreated message in the final receiver need be inherent in this method of transmission, nor is the scheme limited to eight channels. Systems using these same principles could be designed readily for many more channels.

In carrying out the pulse modulation method, the voice wave of each incoming message is sampled 8,000 times a second. Each sampling subsequently results in a pulse of microwave power leaving the transmitting antenna. With no voice input, the pulse associated with a particular channel recurs 8,000 times a second or every 125 microseconds. When the channel is busy, the pulse occurs earlier or later, depending on whether the voice wave is positive or negative. The exact displacement of a pulse from its unmodulated position is linearly proportional to the magnitude of the voice wave at the time of sampling.

The modulation of channel 2 in this manner is illustrated in Figure 5 for the case of a particular sinusoidal input of 1,000 cycles. When the input signal is zero, namely, at points *A*, *E*, and *I* in the figure, the channel pulse will occur at the middle of the position allocated to the channel. Successive diagrams depict the change in the relative position in time of the channel-2 pulses from frame to frame for nine successive frames.

The eight voice circuits connect to the common unit (Figure 6) and go to the transmitting portion of the multiplex which generates marker and channel pulses. These pulses are cabled to the transmitter and are used to turn the transmitter (transmitter-oscillator) on or off for intervals of time corresponding to the length of each pulse. Thus, d-c or video pulses from the transmitting multiplex are translated into pulses of microwave power. These microwave pulses are conducted to the focal point of the transmitting paraboloid where they radiate to illuminate the reflector, the pattern of the reflected radiation being sharply beamed.

Attenuated microwave pulses from the distant transmitter are picked up by the receiving parabolic reflector (Figure 6) and go to the receiver (receiver-converter). The converter changes the microwave pulses to *IF* (intermediate frequency) pulses centered about 58.5 megacycles. After amplification, the *IF* pulses go to the remainder of the radio receiver located on the ground.

Here the *IF* pulses are amplified further, detected; and the resulting video pulses fed to an amplifier. The receiver has automatic tuning so that it is at all

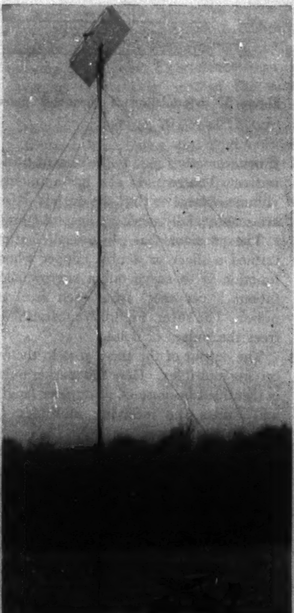

Figure 3. *AN/TRC-6* featuring two-way transmission with a single antenna

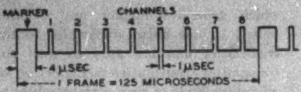

Figure 4. Pulse array at output of transmitting multiplex or input to receiving multiplex

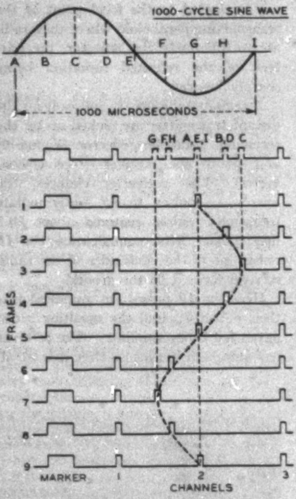

Figure 5. Modulation of channel-2 pulses

times correctly tuned to the distant transmitter. The receiver also has automatic volume control so that the output pulses are substantially constant despite fading.

The receiver has a special feature termed a slicer or double clipper whose function is to carve out a narrow slice (about 5 per cent) from each received pulse. The slicer (Figure 7) substantially frees the pulses of noise.

The output of the slicer goes to the receiving multiplex. Here the marker pulse is identified because of its greater length and causes the receiving multiplex to operate "start–stop."

The sequence of events triggered off by recognition of the marker opens, successively, electronic "gates" to each of the eight channels in turn. After a particular channel gate is opened, the next pulse passes through and is caused to trigger another circuit which starts a direct current flowing. This flow of direct current continues until stopped by the closing of the gate. The starting and stopping of a direct current in this manner generates a length modulated pulse whose trailing edge is fixed but whose leading edge depends upon the exact timing of the received channel pulse. This represents the conversion of pulse position modulation to pulse length modulation. The voice signal is derived by passing the length modulated pulses through a low-pass filter.

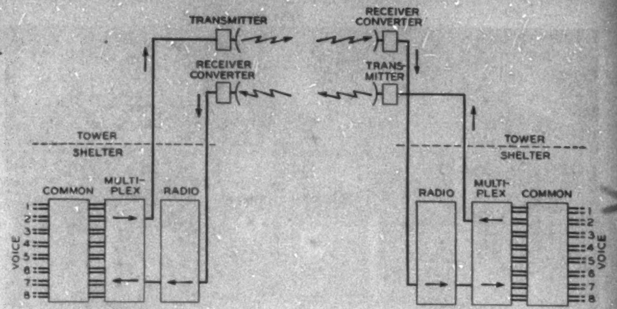

Figure 6. Block diagram of single link system

Specific Embodiment

Method of Operation of the Multiplex

Figure 8 is a block diagram of the transmitting section of the multiplex. An 8,000-cycle oscillator fixes the frame frequency and produces a rectangular wave for starting the marker pulse generator and eight pulse position modulators.

The marker pulses are generated by differentiating the oscillator output and applying the resulting pulses to a tube so operated as to be insensitive to positive pulses. Each negative pulse produces an output pulse which, when limited or clipped, results in a nearly rectangular 4-microsecond pulse at the start of each frame.

Except for channel 1, which will be described separately, short positive pulses are required for starting the pulse position modulators. For channels 7 and 8 it is possible to derive these directly from the oscillator by a differentiating circuit. Exciter circuits, similar in design to the marker generator, start channels 2 to 6. Use of separate exciters for odd and even numbered channels reduces interchannel crossstalk.

Except for channel 1, each position modulator is a multivibrator. This multivibrator is a 2-stage resistance-capacitance coupled amplifier with its output fed back to the input. The circuit is so operated that the first stage normally is cut off and the second conducts. Application of a positive trigger pulse to the grid of the first stage causes it to conduct and cut off the second stage. After an interval of time, depending mainly upon the time constant of the *R-C* circuit which connects the two stages, the multivibrator relaxes, that is, returns to its original condition with the first stage cut off and the second conducting. Each trigger pulse results in an output pulse, the duration of which is determined primarily by the time constant. The time constants are adjusted so that each multivibrator relaxes at the time assigned to that particular unmodulated channel pulse.

Although the time of relaxation of the multivibrator is controlled primarily by its time constant, this time also can be altered by superposing a varying voltage on the grid of the second stage. This grid is the point at which voice frequency voltage is applied from the voice amplifier of that particular channel. Consequently, the trailing edges of the pulses developed by the multivibrator are position modulated by the voice input. A 1-microsecond pulse, produced in a generator common to four channels,

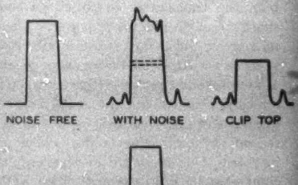

Figure 7. Reduction of noise by slicer or double clipper

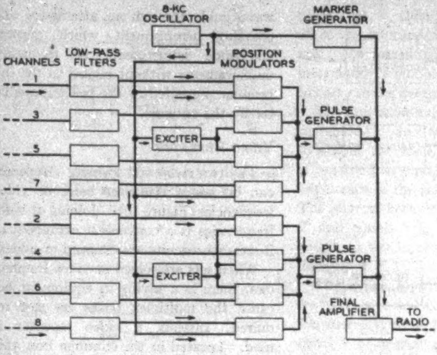

Figure 8. Block diagram of transmitting multiplex

is derived from this trailing edge. The two groups of channel pulses and the marker are combined and go to the radio transmitter.

Because little time exists between the marker and the first channel pulse, channel 1 differs from the other channels in that it does not require a multivibrator to delay the channel pulses. For this channel, therefore, a simpler circuit is possible. This circuit produces a length modulated pulse by clipping a saw-tooth shaped pulse at a voltage determined by the voice sample. The desired channel-1 pulses are derived from the leading edges of these length modulated pulses.

In the receiving section of the multiplex (Figure 9), video pulses from the radio are amplified and the circuit branches into two paths, one of which serves to select the marker pulse. The marker selector is so arranged that its output voltage is proportional to the duration of an input pulse. Marker pulses produce voltage peaks about four times those resulting from channel pulses. The marker amplifier is negatively biased so that the channel pulses do not cause plate current to flow. Therefore only the marker is passed. A new pulse derived from the trailing edge of the selected marker is fed to a multivibrator (called square wave generator) which thus is synchronized accurately with the marker. This symmetrical square wave is used either directly or through intervening sweep generators to operate the gate generating circuits.

To separate the channels, properly timed gate pulses are used. Each gate is a rectangular pulse 13 microseconds long. The gates must be delayed by varying amounts from the edges of the square wave. This is accomplished by generating sweep or saw-tooth voltages and starting the gate pulses at the time the respective sweep voltages cross a reference voltage. The circuit constants for the various channels are different so as to vary the rate of rise of the sweep voltage and thereby locate each gate at its assigned time.

It was noted previously that the received video pulses branch into two paths, one of which selects the marker. The second path includes a 2-stage amplifier which produces new pulses derived from the trailing edges of the received pulses.

These new pulses are fed to the input of the eight channel converters. To each converter is fed also the output of the appropriate gate generator. The gate pulse voltage plus its corresponding channel pulse voltage is appreciably greater than the voltage of the other channel pulses. For example, gate 1 voltage plus channel 1 voltage is greater than channel pulses 2 to 8. Channel 1 is separated from the other channels and this position modulated pulse is converted to a length modulated pulse by a multivibrator so operated that it can be triggered by the combined gate and channel pulse voltages but not by either alone. This circuit not only separates the channels but also converts pulse position modulation to pulse length modulation. When the channel pulse is superposed on the gate pulse, sufficient voltage is obtained to start the multivibrator so that the first stage conducts and remains conducting for the remainder of the gate pulse interval. At the end of the gate pulse, the first stage in the multivibrator circuit is cut off, thereby returning the multivibrator to its original condition. Because the multivibrator is started at a variable time dependent upon the position of the channel pulse and is cut off at a fixed time established by the end of the gate pulse, the output pulses vary in length in accordance with the instantaneous position of the input pulses. By passing these length modulated pulses through a low-pass filter, the audio signal is obtained and all modulation products above the highest audio frequency are eliminated.

Special Radio Features

While two radio frequencies suffice for operation over a single link, a choice of four frequencies is available as a further means of reducing interference between systems or between sets of the same system when repeaters are operated in tandem. Although designed to operate on fixed frequencies, the equipment is capable of being adjusted continuously to operate at any frequency between the extremes. Both transmitter and receiver-converter use small oscillator tubes of the reflex velocity-modulated type

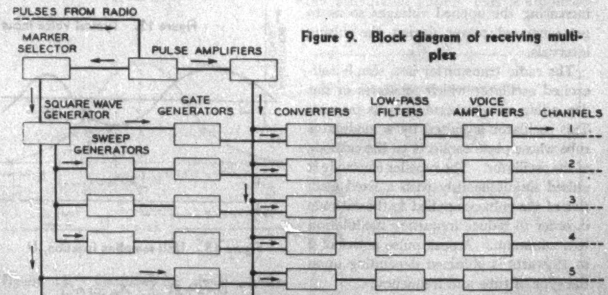

Figure 9. Block diagram of receiving multiplex

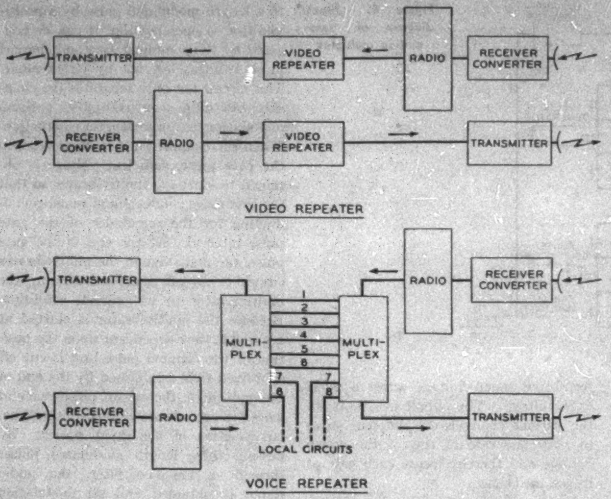

Figure 10. Alternative repeater arrangements

having a flexible cavity which is mechanically varied to get the approximate frequency. The exact frequency is obtained by varying the repeller potential.

Each antenna has a gain of approximately 33 decibels over a simple half-wave antenna. The polarization is normally horizontal but vertical polarization can be used if desired.

A feature of a pulse modulated oscillator is that because power is not generated continuously, the heating of the tube is less than for a continuous wave system and advantage may be taken of this by increasing the applied voltages so as to get increased power during the pulsing intervals.

The radio transmitter is a simple self-excited oscillator which oscillates or not depending on whether a pulse is present. This oscillator is pulsed by a modulator tube whose plate connects to the cathode of the oscillator. The repeller electrode is pulsed simultaneously with a fixed fraction of the voltage applied to the cathode in order to reduce frequency modulation to a minimum. A peak pulse power of 2 to 10 watts is obtained depending upon the type of tube and frequency.

Radio frequency power from the oscillator passes through a short coaxial lead which radiates into a wave guide through an impedance matching arrangement. The wave guide is tuned with a plunger at one end and delivers power to the antenna at the other end. A coupling arrangement is provided on the latest model to permit in-service measurement of microwave power without affecting the power delivered to the antenna. Microwave power is measured with a portable direct-reading power meter.

In the latest model of the receiver-converter, the beating oscillator is fed to a wave guide through an attenuator and coupling arrangement which permits amplitude adjustments or frequency measurements without regard to the impedance presented to the beating oscillator by the antenna.

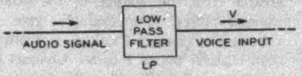

Figure 11

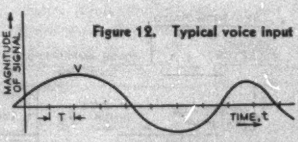

Figure 12. Typical voice input

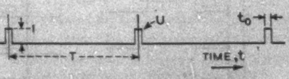

Figure 13. Unit sampling function, U

$$U=k+2k\sum_{m=1}^{\infty} A_m \cos mCt$$

$$f_c=\frac{C}{2\pi} \qquad \frac{1}{f_c}=T$$

$$k=\frac{t_0}{T} \qquad A_m=\frac{\sin mk\pi}{mk\pi}$$

Video Repeater

The two radio sets forming a repeater can be joined either at voice or video frequencies (Figure 10). Joining at voice frequencies is a convenient arrangement if message circuits are dropped or added.

When sets are joined at video frequencies, there is a saving in equipment because the multiplex boxes are not required. Instead, a video repeater is used. Located in the common box, this repeater provides order wire facilities for monitoring and modulating channel 1 so that a relay attendant will be able to signal and talk to any other station along the route. In addition, the received pulses are lengthened to compensate for their having been shortened, chiefly by the preceding transmitter.

Special Equipment Features

The mechanical and equipment features of *AN/TRC*-6 were governed largely by Signal Corps specifications. Important objectives such as ruggedness, portability, maintenance, storm-proofing, ease of setup, weight, and dependability were given proper balance to obtain a set suitable for military use.

The 50-foot sectionalized welded aluminum tubing tower consists of six 8-foot sections from two to three feet square with provision for omitting sections to obtain lesser heights. The four intermediate sections are tapered and may be telescoped together for shipping. The top section rests in the base section together with the hoisting boom during shipment.

With the tower properly guyed, the movement of the antenna beam pattern is restricted to a fraction of a degree even in a 60-mile wind. The sets have oper-

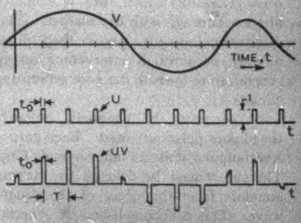

Figure 14. Diagram of V and U

UV is the result of sampling *V*

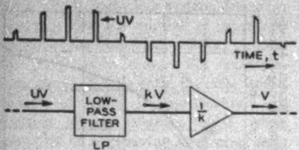

Figure 15. Passing *UV* through a low-pass filter and amplifier to obtain *V*

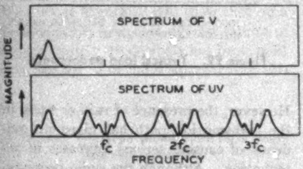

Figure 16. Spectrum analysis of *V* and *UV*

ated successfully through two hurricanes. The electrical and mechanical performance have proved satisfactory in actual field usage even when the antennas, top beam section, platform, tower, and guys were coated with more than three-quarters of an inch of ice.

Pulse Modulation Principles

The discussion to follow is a treatment of pulse modulation including the sampling principle and time division. To enumerate important features:

1. It is the sampling principle which specifies the least number of points required to reproduce a changing voice wave.

2. It is the job of the modulated pulses to carry this information to the distant receiver.

3. Pulse position modulation, like frequency modulation, reduces noise at the expense of band width.

4. A method of scaling is described whereby time division is regarded as a scale model of a single channel system.

In carrying out the pulse modulation method, the voice signal usually is passed first through a low-pass filter (Figure 11) to cut out all frequencies above an upper limit which is less than one-half the sampling frequency, f_c. After filtering, the signal is designated V and termed voice input.

V (Figure 12) is the instantaneous value of a short portion of a voice signal which might have passed the low-pass filter of Figure 11. The sampling principle teaches that if V is sampled instantaneously at regular intervals T, where T is less than a half period of the highest frequency in V, then the samples contain all of the information of the original signal. The magnitude of a sample is the information a modulated pulse is required to carry to the distant receiver.

The unit sampling wave (Figure 13) is designated U and will be used presently to sample the voice. Mathematically U is regarded as equal to a d-c term plus harmonics of the sampling frequency, f_c. The interval between pulses is designated T and equals $1/f_c$. The ratio of pulse length to the interval between pulses is designated k.

The voice input and unit sampling wave are repeated at the top of Figure 14. Because U is either one or zero, the product UV is an analytical process for sampling the voice. The result is a series of positive and negative pulses. When U is one, the product is equal to the voice input and at all other times zero.

Because UV is similar in appearance to amplitude modulated pulses (Figure 15), it is not surprising to find that an attenuated replica of V is obtained simply by passing UV through a low-pass filter. In the process of sampling and filtering, V is reduced by a factor k. Amplifying by a factor $1/k$ restores V to its original value.

To bring out more concretely that passing UV through a low-pass filter produces the original signal, Figure 16 presents a spectrum analysis of V and also UV. The upper diagram is the spectrum of V. After sampling V becomes UV. The spectrum of UV is the spectrum of V, small but exact, together with upper and lower sidebands about harmonics of f_c.

Remembering UV is the result of sampling, we can draw a conclusion; if, in a communication system, the resultant sampling UV is transmitted to the distant receiver either by direct transmission or by conversion to pulse position modulation, the voice signal can be derived simply by passing through a low-pass filter followed by an appropriate amplifier.

The simple physical concepts of pulse position modulation (Figure 17) will be reviewed by considering an idealized 1-channel system. With no voice input, pulses of fixed amplitude and length recur at regular intervals T. When the channel is busy, each pulse is displaced from its unmodulated position and occurs earlier or later depending on whether the voice signal is positive or negative. The exact displacement is linearly proportional to the instantaneous value of the voice wave at the time of sampling. With full input, the maximum excursion, neglecting the length of the pulse, is plus or minus $T/2$ or $\pm 1/2\,f_c$.

Figure 7 illustrates the reduction of noise by the slicer or double clipper. A practically noise-free pulse is delivered by the transmitting multiplex. A long radio path attenuates the pulse and noise is introduced by the receiver, producing a typical pulse with noise as indicated. The double clipper works in two steps. First, the top of the wave is shaved off reducing noise superimposed on the pulse. Second, the bottom of the wave is chopped off leaving a nearly noise free but attenuated pulse. This is amplified

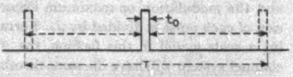

Figure 17. Single-channel pulse-position-modulation system

$$\pm\frac{T}{2}=\pm\frac{1}{2f_c}$$

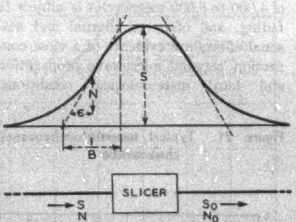

Figure 18. Equation for signal-to-noise improvement resulting from slicer

$$\frac{S}{N}=\frac{1}{B\epsilon}\qquad\frac{1}{\epsilon}=\frac{SB}{N}$$

$$\frac{S_0}{N_0}=\frac{1}{2\sqrt{2}\epsilon f_c}\qquad\frac{S_0}{N_0}=\frac{S}{N}\times\frac{B}{2\sqrt{2}f_c}$$

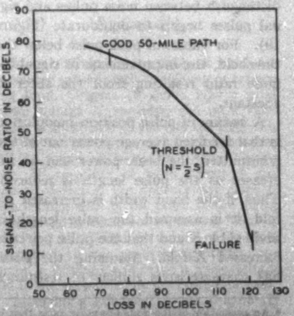

Figure 19. Decibel signal-to-noise ratio versus path loss

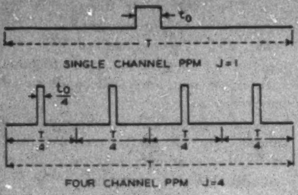

Figure 20. Formation of *J* channels by time division

and delivered to the receiving multiplex. It should be noted that whereas the effects of noise on amplitude can be practically eliminated by the slicer, nevertheless, the effects of noise on the exact timing of the pulse remain. However, the magnitude of this error in timing can be reduced by increasing the band width.

Because the exact time of occurrence of these nearly noise-free pulses is changed by the presence of noise at the input to the slicer, this effect, namely, the variation in exact time of occurrence of the pulses, will register as additional noise in the final receiver. This additional noise is evaluated by an approximate formula (Figure 18) for the signal-to-noise improvement due to the slicer (see Appendix). $B/2\sqrt{2f_e}$ is the improvement where B is the *IF* band.* This is the improvement for a single channel system whose maximum excursion approaches plus or minus $T/2$.

Because of the slicer, noise affecting the remainder of the system is restricted substantially to that which occurs during the time that the pulse amplitude is within the range of the slicer, provided the peak amplitude of the noise is less than one-half the peak amplitude of the signal. As the noise exceeds this limiting value or threshold, the ability of the receiver to distinguish between noise pulses and signal pulses begins to deteriorate (Figure 19). For values of peak noise below the threshold, the improvement in signal-to-noise ratio resulting from the slicer is constant.

A feature of pulse position modulation is that for fixed average power out of the transmitter, the peak power can be increased as the pulse length is reduced. Thus, if the band width is increased K-fold, it is assumed the pulse length is divided by K and that the pulse power is increased K-fold. Assuming the noise has characteristics similar to resistance noise, the signal-to-noise ratio at the input to the slicer and, therefore, the threshold, are independent of band width. Because of the K-fold wider band, the maximum rate of rise and fall of the pulse will be increased K-fold, and consequently the signal-to-noise ratio at the output of the slicer further is increased K-fold which corresponds to a further improvement in decibels of $20 \log_{10} K$. This is in contrast to a conventional frequency modulation system where, over a path of specified attentuation, the threshold represents a limitation on the signal-to-noise ratio obtainable with a given amount of carrier power.

Passing to the theoretical concept of time division, a time division multiplex can be treated by first considering an idealized 1-channel system. Assuming one channel and maximum excursion of the pulse, the band width will depend upon the signal-to-noise improvement sought. For double sideband transmission, the *IF* band will be twice the video. The band width of the transmitter should equal that of the receiver. The pulse should be as short as the band width of the system will tolerate.

To provide J channels by time division (Figure 20), all significant band widths are increased J-fold; the length of each pulse is divided by J; the peak power of each pulse is multiplied by J; J pulses appear in each frame; the average power of the transmitter is increased J-fold; and the modulation or maximum excursion of each pulse is divided by J. Forming a scale model in this fashion, the J-channel system will have the same threshold and same signal-to-noise ratio as the idealized 1-channel system.

Transmission Performance

Microwave propagation at frequencies of 4,000 to 5,000 megacycles is subject to fading, and observed diurnal and seasonal effects are evidence of a close connection between microwave propagation and local meteorological conditions. However, the presence of rain or snow in the transmission path has given no evidence of causing serious increases in attenuation. Although the radio propagation may vary by very substantial amounts, the transmission stability of the message circuits is extremely good.

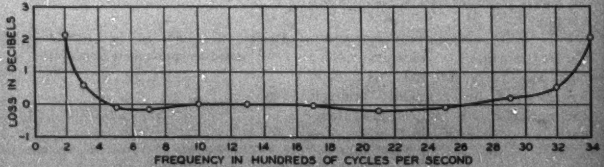

Figure 21. Typical transmission–frequency characteristic

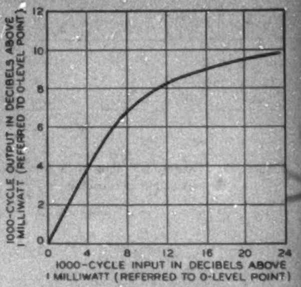

Figure 22. Typical load performance

The eight message circuits which each *AN/TRC*-6 system provides are high quality telephone circuits and meet commercial standards for long distance telephone transmission. The individual message circuits may be terminated 2-wire or 4-wire and may be used to transmit signaling, dialing, facsimile, pictures, or multichannel voice frequency telegraph. In the latter instance, a single message circuit will handle as many as 18 separate teletypewriter facilities.

As indicated by Figure 21, the transmission–frequency characteristic of each of the eight channels varies less than one decibel from 300 to 3,300 cycles, and, as shown by Figure 22, each channel is capable of handling inputs up to about plus eight decibels above a milliwatt at the zero level point (2-wire input) without appreciable overloading. The load capacity of the receiving multiplex is sufficient so that when terminated 4-wire, the circuit can be adjusted to nine decibels gain. Receiving potentiometers provide a continuous adjustment of channel gain over a 15-decibel range.

Inspection of Figure 23 shows that the circuit noise observed on a single link is dependent upon the attenuation in the path. For a many link system, this noise increases with the number of repeaters. However, the individual message circuits are particularly quiet. For example, over a good 50-mile path, the noise in the absence of fading is approximately six decibels above reference noise when measured at a receiving level nine decibels below zero level. In terms of rms signal-to-noise ratio, assuming the signal is plus eight decibels above a milliwatt at zero level, this corresponds to a signal-to-noise ratio of 75 decibels. Further inspection of Figure 23 will show that on a 50-mile path, a margin against failure of approximately 48 decibels has been built into the set, and fades approaching this magnitude can be encountered before the system fails.

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* An evaluation of the signal-to-noise ratio that is similar in principle is included in an article entitled "Pulse Time Modulation" by E. M. Deloraine and E. Labin in Electrical Communication, volume 22, 1944, number 2, pages 91–8.

The intervals of time that are significant in determining the exact position in time of a channel pulse are small. For example, in Figure 23, if the relative random jitter in the trailing edge of the channel pulses had been as great as one thousandth of a microsecond, the noise meter readings for sufficiently low values of path attenuation would have been slightly more than 3 decibels higher than the experimental values observed.

Two-way voice transmission over radio links totalling 1,600 miles and one-way over 3,200 miles have been accomplished successfully in demonstrations. To do this, ten radio sets were set up to form five two-way 8-channel systems each operating simultaneously over the same 40-mile air path. The five links were connected in tandem to form a two-way 8-channel system 200 miles long with four intermediate repeaters. By connecting all channels in tandem a two-way 1,600-mile circuit was obtained. Finally, by connecting both directions together, excellent 1-way transmission was obtained over an air path 3,200 miles long. To illustrate the over-all stability of the 3,200-mile circuit, Figure 24 is a typical 24-hour recording of the received level of a 1,000-cycle test tone and it can be seen that the maximum variation is less than plus or minus 0.5 decibel.

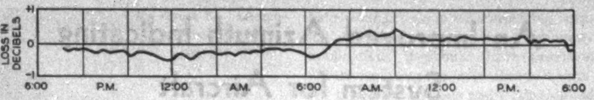
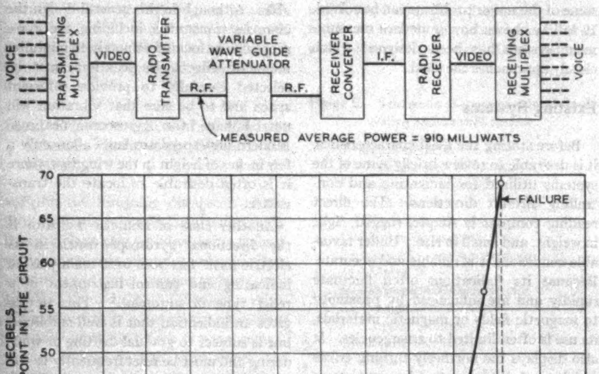


Figure 24. Transmission stability of 3,200-mile circuit

Figure 23. Message circuit noise for a single link versus path loss



Conclusion

Although this equipment was designed for military use, its basic principles and design features can be used to provide telephone and other communications to the public. Sharply beamed radiation and reception, pulse modulation and time division multiplex are all factors which tend to make such systems attractive and their economic field in competition with other alternatives needs to be determined.

Appendix

Noise will modify the times at which the incoming pulses reach the slicing or triggering level. Figure 18 shows how a small noise voltage N, superimposed on a video pulse of amplitude S, changes the triggering time by an amount ϵ. From the geometry of the figure, $S/N=1/B\epsilon$. After passing the IF amplifier of band width B and second detector, the approximate slope of the leading edge of the pulse at the slicing level is SB and the corresponding "time of rise" is $1/B$. When the output of the slicer is passed to the receiving multiplex, the error in timing ϵ will cause a noise voltage N_0 in the output of the receiving multiplex.

Assuming a 1-channel system (Figure 17) and supposing that the maximum excursion of a pulse can assume its limiting value of plus or minus $1/2f_c$, plus or minus $1/2f_s$

An Improved Azimuth Indicating System for Aircraft

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Synopsis: Because of the violent maneuvering possibilities in modern aircraft, the usefulness of azimuth indicators in avigation depends on circumstances not existing in marine navigational applications. New problems in direction indicating instruments therefore have been introduced, such as the perplexing problem of obtaining indication after loops or rolls. Azimuth indicating systems now used in aircraft have not given a completely satisfactory solution to such problems.

This paper summarizes the progress from the direct reading compasses to monitored gyroscopes and concentrates on the unique problems in aircraft where space limitations, reliability, and operation at all altitudes are important. It indicates how azimuth indicators and auto pilot controls close to the ideal are obtainable by the use of special features including a novel "nontumbling" mechanism and an electromagnetic type of gimbal adjusting device.

DURING the past few years more emphasis has been placed on the development of a directional indicating and control system which will meet the new rigid requirements for avigation as well as the older well-established needs of navigation which are still used in guiding aircraft. The first systems for use in aircraft followed the natural development trend towards improvement of navigational compasses. However, the rotational and translational accelerations which are experienced in modern aircraft bring forth new and, until recently, unsolved problems in azimuth indication. A major and perplexing problem is that of obtaining indications during steep banks, rolls, and loops. It is the purpose of this paper to point out the principal features needed in an azimuth system for use in modern aircraft and to indicate how some of the major problems can be solved. It will be shown how a modern measurement system has been designed which closely approaches the ideal.

Existing Systems

Before stating the ideal characteristics, it is desirable to review briefly some of the systems utilized for indicating and controlling aircraft directions. The direct reading compass is simple, rugged, light in weight, and small in size. Under favorable conditions it is reliable and accurate. Because its indications often fluctuate rapidly and are influenced by proximity to magnetic fields or magnetic materials, its use is often limited to emergencies. It also displays the northerly turning errors which are well known to pilots and others familiar with compasses.[1] The remote indicating compass helps to solve the difficulty resulting from magnetic fields and materials by locating the direction sensitive element in a position where such influences are minimized. However, it does not solve the problems of northerly turning errors or pointer oscillation.

In certain cases the stabilization of the magnetic compass has been found desirable[2] in order to eliminate the effect of the vertical component of the earth's field which is responsible for the northerly turning errors. This is done in the gyro-stabilized compass by mounting the direction sensitive element on a vertical axis gyroscope. The element thus is maintained in a level position during maneuvers and while flying in rough air. However, because the vertical gyroscope is kept vertical by gravity control, it will have some errors during maneuvers of long duration since centrifugal forces influence the apparent direction of the gravitational force. It also has the disadvantage of heavy weight and large size in the forms that are generally available. Although in this general design the compass transmitter including the gyroscope can be located remotely to minimize magnetic effects, its location must be selected carefully to provide sufficient space and to be sure that vibration will not damage the gyroscope bearings. Modern high speed airplanes allow only a few inches of height in the wing tips where it is often desirable to locate the transmitter.

Another class of azimuth indicator is the directional gyroscope which in its electric form[3] has been used mainly as an indicating and control instrument in a relief type of autopilot.[4] This device gives an indication that is well-stabilized but is subject to gradual drifting or wandering and must be reset frequently to the magnetic compass heading. A part of the drift may be the result of gimbal bearing friction, unbalance, or mechanical imperfections, but a more fundamental cause is the rotation of the earth which produces an apparent drift depending on the geographical location and heading of the directional gyroscope. Resetting is inconvenient and may result in an error if the compass happens to be under the influence of temporary accelerations when the heading is transferred from the compass to the gyroscope.

To avoid these difficulties the directional gyroscope can be reset automatically and continuously to agree with the compass.[5] Such an arrangement utilizes a remote compass transmitter and thus minimizes magnetic difficulties. Because it utilizes a directional gyroscope, its indication is stable. However, there still remains an unsolved problem in conventional directional gyroscopes, whether compass controlled or not. This is the

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represents the peak modulation (displacement in time) of a pulse from its unmodulated position. The corresponding rms displacement when the pulses are modulated fully by a single frequency test tone approaches $1/2\sqrt{2}f_c$. If S_0 denotes the corresponding rms signal voltage in the output of the multiplex, $S_0=K/2\sqrt{2}f_c$ where K is a constant factor of proportionality.

In Figure 18, N can be regarded as the voltage of an instantaneous noise sample taken at the time of triggering and ϵ as the corresponding error in timing caused by the noise sample. N can vary from sample to sample. The approximate rms value of ϵ can be taken as the rms value of N divided by SB. In this case N_0 will denote the rms noise voltage in the output of the multiplex. If the noise voltage at the input to the slicer has the characteristics of resistance noise, its rms value will equal the rms value of the instantaneous noise samples N.

Since

$$S_0=K/2\sqrt{2}f_c$$

and

$$N_0=KN/SB$$

then

$$\frac{S_0}{N_0}=\frac{S}{N}\times\frac{B}{2\sqrt{2}f_c}$$

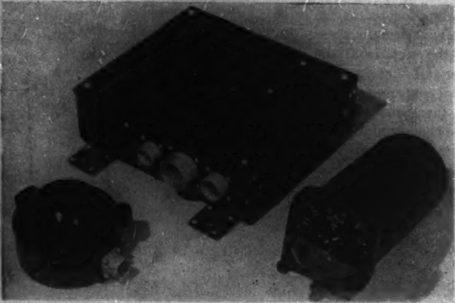


Figure 1. Components of azimuth indicating system

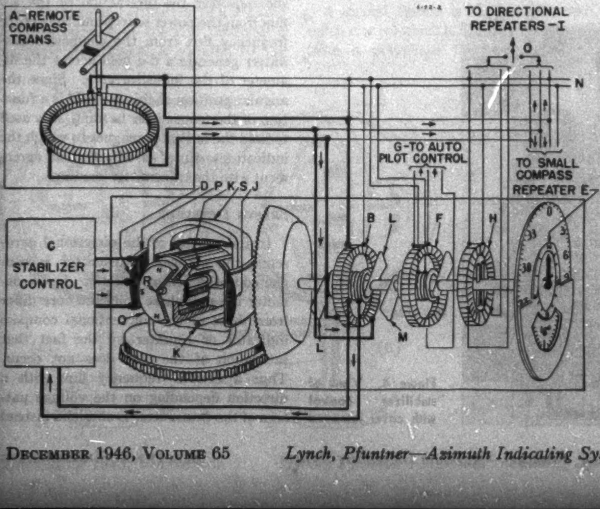
elimination of the error in indication or complete loss of indication caused by the upsetting of the gyroscope motor because of insufficient freedom of movement within its gimbal mounting. With modern aircraft violent maneuvers are possible and loops or rolls must be expected with military planes. These maneuvers will cause normal directional gyroscopes to upset.

Ideal Characteristics of a Modern Azimuth Indicating System for Aircraft

From the foregoing discussion of the limiting features in existing arrangements, it can be seen that an ideal azimuth indicator or control system for use on modern aircraft must meet certain requirements to avoid the difficulties described. Without attempting to cover all the requisites,[6] one may state that such a measurement system should be light and small, and should occupy minimum panel space. Such a system should have provision for minimizing the effects of magnetic fields or magnetic materials, and its accuracy should meet the requirements of the application. It should produce an indication which is stable. Its reading should not be excessively disturbed by normal or even violent maneuvers. It also should have some provision for minimizing turning and acceleration errors. For many applications it should have provision for transmitting the indication to a number of remote stations and it also should provide the necessary signals for azimuth stabilization or azimuth autopilot control. An additional feature considered of much importance is an emergency compass, so that if the gyroscope fails there will still remain a readable indication which is independent of local magnetic effects. A further very important feature is the characteristic of "safe failure" if anything should go wrong with the system.

Figure 2. Schematic diagram of azimuth indicating and control system

B—Selsyn detector
D—Torque motor
F—Auto pilot "pick-off"
H—Subtransmitter
J—Winding
K—Alnico bar magnet
L—Magnetic vane mounted on shaft
M—Vanes
N—115-volt 400-cycle source
O—Switch
P—Copper cylinder
Q—4-pole disk magnet
R—Motor shaft
S—Main gimbal



Important Features for an Ideal System

Three important features which help to meet these requirements are compass control, automatic gimbal adjustment, and nontumbling action. The first tends to hold the gyroscopic indication in agreement with the average compass position regardless of any natural tendency to drift. The second continuously tends to adjust the gimbals to a position at right angles to each other so that they will not drift into line and cause an upsetting condition. The third feature makes it possible to operate satisfactorily even during violent maneuvers by eliminating the undesirable tumbling action which occurs in normal limited angle directional gyroscopes.

General Description of an Improved System

The following will show how an azimuth control system, which very closely approaches the requirements of the ideal, has been obtained by the controlling or monitoring of a small nontumbling directional gyroscope by means of a remote magnetic compass. The basic devices required are shown in Figure 1 and will be discussed in detail under "Operating Principles."

In order to overcome the tendency of the gyroscope to drift in azimuth, a magnetic compass transmitter of the remote indicating type corrects the heading of the gyroscope continuously, but at a very slow rate. By this means both the natural tendency of the gyroscope to wander and the large oscillations of the compass are reduced to negligible values.

A second harmonic signal from the remote compass transmitter *A*, as shown in Figure 2, is applied to a Selsyn detector *B* located on the azimuth axis of the gyroscope. Any angular difference between the compass and gyroscope is translated

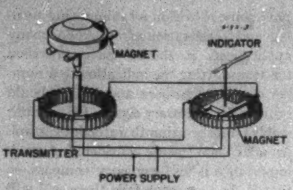


Figure 3. Basic diagram of Selsyn repeater circuits

into a voltage which is amplified in a stabilizing control box *C* and used to correct the gyroscope heading. The control is accomplished in a torque motor *D* attached to the gyroscope.

The headings of the compass transmitter and the gyroscope instrument at any instant may be compared by means of a small compass repeater *E* located in the face of the directional gyroscope. This repeater, which is connected directly to the remote compass transmitter, provides both a check on the correct functioning of the system and a very important stand-by instrument in case of a failure of the gyroscope or automatic control.

An inductive type polyphase electromagnetic "pick-off" *F* generates alternating voltages *G* capable of azimuth control of an automatic pilot or other equipment.

Because many airplanes require several simultaneous indications of direction in various parts of the airplane, a subtransmitter *H*, as shown in Figure 2, which is similar in principle to the remote indicating compass transmitter, also is provided. This is capable of operating a number of light compass repeaters *I* located as desired about the aircraft. Normally these repeaters duplicate the signal of the directional gyroscope and therefore are not subject to the usual compass errors or oscillations. Should the gyroscope fail, however, they may be connected directly to the compass transmitter by switch *O* shown in Figure 2. With this connection they provide a conventional remote compass system.

Operating Principles

A system such as the one described has many features which are responsible for its satisfactory operation. Included in a list of such features are

1. The Selsyn detector to determine the difference between compass and gyroscopic indications.
2. The stabilizer control unit that detects which direction the torque motor should run and controls that motor.
3. The torque motor for correcting the gyroscopic reading.
4. The gyroscope motor.
5. The repeater circuits which make possible the transmission of the stabilized gyroscopic indication to a large number of directional repeaters.

The operating principle of units furnishing these features will be outlined before discussing the details of the very important automatic gimbal adjustment and nontumbling action features.

REMOTE COMPASS SYSTEM

The compass transmitter and repeater usually are connected in a circuit similar to that shown in Figure 3. Double frequency voltages are generated in the transmitter windings by the interaction of the d-c flux from the north-seeking compass magnets and the a-c flux from the power source in a saturatable core material. The generation of double frequency voltages by this saturating means and their use in repeater instruments already has been described adequately.[7]

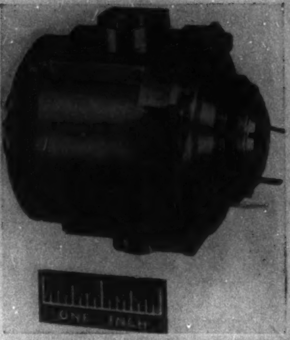


Figure 5. Gyroscope motor for compass controlled directional gyroscope

The distribution of the voltages about the core is determined by the angular position of the transmitter magnets relative to the core and coils, and since the magnets are free to align themselves with the earth's magnetic field, double frequency voltages are obtained whose relative magnitudes are a function of magnetic heading of the aircraft. The usual practice is to impress these voltages upon an indicator whose magnetic construction is similar to that of the transmitter. In the indicator the interaction of the a-c flux from the power source and the double frequency flux from the compass transmitter generate a d-c field across the diameter of the indicator core. Since the angular position of this d-c field is a function of the transmitter heading, it is used to hold the indicator magnet to which the indicating pointer is attached in agreement with the transmitter heading.

SELSYN DETECTOR

In the control of the directional gyroscope the compass transmitter is connected to the Selsyn detector. In the detector the polyphase coils and core differ from those of the conventional compass indicator or repeater by the fact that saturation of the core does not occur. Thus a double frequency flux with a direction depending on the voltage pattern of the transmitter is obtained instead

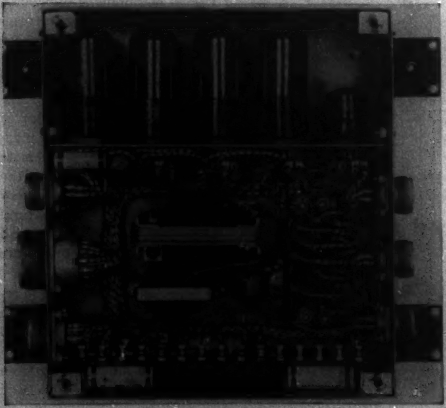


Figure 4. View of stabilizer control with cover removed

of a d-c field. Two magnetic vanes mounted on a shaft, as shown by *L* in Figure 2, can pick up this flux and guide it through a detector coil to detect the angular position of this double frequency flux. If the vanes are exactly in line with the a-c field, a maximum voltage will be induced in the coil, but if the vanes are turned 90 degrees from this position, zero voltage will be induced. By mounting the vanes on the gyroscope azimuth shaft the zero voltage position is used to detect when the gyroscope azimuth axis is in agreement with the compass heading. Any voltage present in the detector coil indicates an error in the gyroscope heading, the direction of which is determined by the polarity (in phase or 180 degrees out of phase from a reference voltage) of the alternating voltage.

STABILIZER CONTROL UNIT

The detector output is impressed upon an electronic control unit which amplifies and rectifies the signal to give a d-c output for operation of the torque motor to bring the gyroscope to the correct heading. A photograph of the stabilizer control unit with the cover removed is shown in Figure 4. The signal first is amplified in a conventional 2-stage amplifier circuit. It then is rectified. During rectification it is combined with a properly phased master voltage to obtain a direct current which reverses if the polarity of the input reverses. This direct current in turn is amplified and used to control the torque motor.

TORQUE MOTOR

The torque motor *D*, Figure 2, is operated by direct current from the stabilizer control unit circulating between the center tap and either end of a winding *J*. The current in the coil produces a magnetic field inside it parallel to the gyroscope main gimbal axis and in a direction determined by the direction of the correction required. Four alnico bar magnets *K* are placed on the sides of the motor frame with their axes parallel to the motor shaft. The magnetic field from the coil reacts with these magnets to produce a torque in such a direction that precession will take place about the azimuth axis. Precession will continue until the compass and gyroscopic indications are in agreement.

GYROSCOPE MOTOR

The gyroscope motor shown in Figure 5 is a 3-phase 2-pole hysteresis-type motor which operates synchronously at 24,000 rpm and is capable of furnishing a powerful starting torque. Adequate moment of momentum is provided by means of an external rotor shell of a tungsten alloy, and this together with the high rotating speed gives the necessary stabilizing torque. Special bearings designed to withstand high vibration accelerations give these motors improved characteristics.

REPEATER SYSTEMS

In order to utilize the stabilized directional gyroscope for other purposes than as an individual indicator, additional repeater systems are employed. One of these is the subtransmitter *H* shown in Figure 2. In principle this transmitter is equivalent to that of the remote compass except that the transmitter magnet is attached to the gyroscope azimuth axis. By means of a circuit similar to that shown in Figure 3, the heavily damped gyroscope subtransmitter signal may be used to actuate a number of standard remote directional repeaters located conveniently about the aircraft.

A second system is the autopilot azimuth control which is initiated by a signal from a polyphase "pick-off" mounted around the gyroscope azimuth shaft as shown at *F* of Figure 2. The construction of this "pick-off" is similar to that of the Selsyn detector, but its operation is somewhat different. The power supply frequency is supplied to the single center coil and the vanes *M* distribute the flux to the outer polyphase coils in accordance with the angular heading of the gyroscope. Voltages produced in the polyphase coil are therefore a function of the gyroscope heading. These voltages may be used for accurately controlling and stabilizing various directional devices.

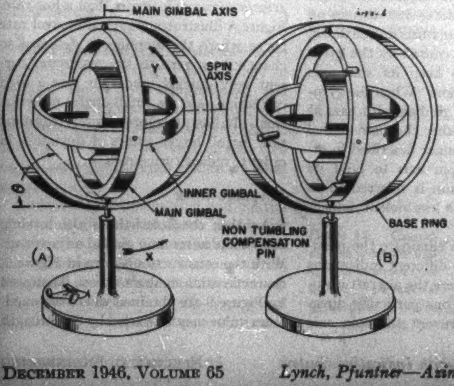


Figure 6. Typical toy gyroscope

A. Standard type

B. Special type equipped with nontumbling compensation

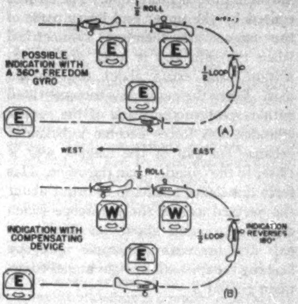

½ ROLL
POSSIBLE INDICATION WITH A 360° FREEDOM GYRO
E
½ LOOP
(A)
WEST
EAST
½ ROLL
W
INDICATION WITH COMPENSATING DEVICE
½ LOOP
INDICATION REVERSES 180°
(B)

Figure 7. Indication during an Immelman turn

A. Without a compensating device

B. With a compensating device

Automatic Gimbal Adjustment

A very important feature is used in this compass-controlled gyroscope system. An automatic method of keeping the gimbals at right angles to each other prevents the horizontal gimbal from precessing into line with the main gimbal. Such a lineup of the gimbals causes the gyroscope to upset. Figure 2 shows the construction schematically. A copper cylinder *P* is attached to the instrument frame in such a manner that it surrounds the gyroscope motor and its axis coincides with the main gimbal axis. In the actual design illustrated, this cylinder also serves as the mechanical support for the torque motor coil. A 4-pole disk magnet *Q* is attached to the end of the motor shaft *R* and turns with the motor rotor at 24,000 rpm. Eddy currents are induced in the copper ring by the spinning magnet and produce an electromagnetic drag acting on the motor rotor.

As it revolves the upper part of the magnet moves in one direction with respect to the cylinder, and the lower part moves in the other direction. The torques caused by the upper and lower parts of the magnet are, therefore, symmetrical when the spin axis is level, and no resultant drag exists about the vertical axis. When the motor axis becomes tilted with respect to the plane of the copper cylinder, the forces become unbalanced because one-half of the magnet disk is closer to the cylinder than the other. This force unbalance results in a torque about the vertical axis of the gyroscope which causes precession about the horizontal axis and prevents gyroscopic upset by keeping the spin axis at right angles to the main gimbal *S*.

Nontumbling Gimbal Reversal Feature

It previously has been pointed out that the system should indicate correctly after all the maneuvers that an aircraft may be expected to perform. The maneuvers include loops and rolls which cause conventional directional gyroscopes to upset because the angular movement of the spin axis in the main gimbal is limited. Universal freedom mounting can be designed to reduce the hazard of upset. A toy gyroscope such as shown in Figure 6A illustrates the general construction. Problems arise, however, which make it impractical or, at least, undesirable.

One major problem is illustrated by the fact that with 360 degrees freedom it is possible to go through certain maneuvers such as an Immelman turn where, after coming out of the maneuver, the instrument will read incorrectly by 180 degrees as shown in Figure 7A. This is further complicated by the fact that when going through a similar maneuver at a slightly different heading, this reversal will not be present.

In order to clarify this principle, consider Figure 6 which illustrates schematically a directional gyroscope. There are two conditions limiting the directional gyroscope's proper use.

1. The spin axis is fixed in direction.
2. The base is fastened to the aircraft and is upright in normal flight.

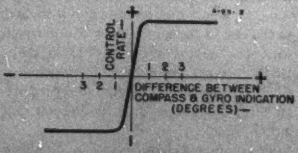


Figure 8. Typical control rate for gyroscope azimuth precession

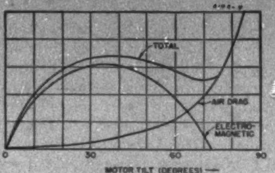


Figure 9. Typical control rate for leveling gyroscope gimbal

The angle θ, Figure 6A, between the plane of the base ring and the plane of the main gimbal is a measure of the gyroscopic indication. It is evident that with only these two limitations, the angle θ can have two values differing by 180 degrees for a particular heading of the gyroscope base ring, since besides the position illustrated, we can turn the spin axis as shown by arrow *Y*, continuing it through the main gimbal, and then we can bring the black end of the spin axis back to the left hand side by rotating it about the main gimbal axis. Thus for the same heading of the aircraft, we have two azimuth readings differing by 180 degrees.

The fact that both of the foregoing readings can exist is verified during an Immelman turn, which is illustrated by the following experiments. If the aircraft is considered heading in the direction shown on the gyroscope base in Figure 6A, and if the base is rotated as shown by the arrow *X* (keeping the spin axis in the plane of the base ring) until the base is upside down, the angle θ will not change. If the base then is returned to normal by rotating about the spin axis, θ still will be unchanged.

If the experiment is repeated in a similar manner, but if the spin axis intentionally is kept slightly off the plane of the base ring, then a phenomenon occurs which at first thought may seem strange. The main gimbal rotates 180 degrees on its own axis as the base turns upside down. This phenomenon is caused by the gyroscope's tendency to keep its spin axis pointed in one direction. When the gyroscope is mounted in gimbals with axes at right angles to each other, the main gimbal must turn in the manner described in order to allow the spin axis to hold its direction. This action is referred to as gimbal error action in less violent maneuvers.

The problem is to eliminate the possibility of having two different indications for the condition where the aircraft is upright and headed in one particular direction. A stop to prevent the spin axis passing through the main gimbal prevents the possibility of two such indications, and the arrangement is schematically shown in Figure 6B. During a loop, when the pin hits the gimbal, it is engaged only momentarily since gyroscopic action turns the main gimbal to free the pin, and therefore no undesirable tumbling action takes place.

The stop is placed so that the inner gimbal cannot quite line up with the main gimbal, and two additional desirable results are attained.

1. Since the inner gimbal will not rotate 360 degrees, spirals can be used to lead current into the motor rather than slip rings. This eliminates undesirable friction.
2. Since the axis of the inner and main gimbal never can line up exactly, the maximum undesirable effects of "gimbal lock" are eliminated. "Gimbal lock" is the condition where the spin axis lines up with the main gimbal axis, and in this position friction of the motor bearings will tend to turn the main gimbal and thus start the directional gyroscope indicator dial spinning.

Operating Characteristics

There are certain operating characteristics of especial interest in the compass-controlled directional gyroscope. Some of these may be called design characteristics since they are easily adjustable in design. These would include the precession rate of the gyroscope azimuth correction, and the leveling rate of the gyroscope gimbal automatic adjustment.

PRECESSION RATE

The advantage of a high rate for correcting the gyroscope indication lies in the faster return of the gyroscope when large errors occur, and also in a reduction in errors caused by the drift of the gyroscope that is being compensated for, as shown in the appendix. However, because the compass may have large errors during turns, a high rate might cause greater gyroscopic errors than a low rate. Figure 8 illustrates a typical control rate curve where the amplifier is designed to change from zero to the maximum current very quickly, and thus give a high correction rate for small angles of deviation. It has an approximately constant rate for all larger angles of deviation in order to reduce turn errors.

LEVELING RATE

Another characteristic is the leveling rate of the automatic gimbal adjustment. With the construction shown in Figure 2, characteristics similar to those illustrated by Figure 9 are obtained where the maximum value may be varied by the strength

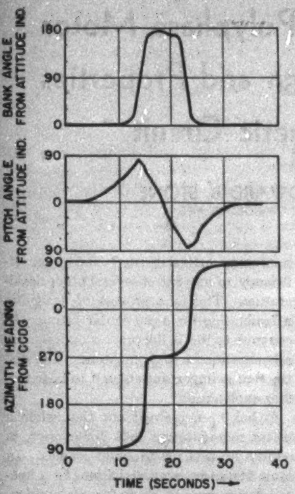


Figure 10. Data typical of actual results during a complete loop

of the alnico bar magnet. Since the dip of the motor normally is not transmitted to the dial, the shape of this curve is not particularly critical. Ordinarily there is nothing to upset the motor to a position outside the range of the copper ring, but if considered necessary, air drag on the motor rotor may be used as a leveling means for the high angles.

Flight Data During Violent Maneuvers

Very important operating characteristics are shown by actual flight data. Figure 10 shows the data taken during a loop starting on a 90-degree heading in a typical flight test.

Conclusions

This system will give satisfactory overall operation covering a wide range of voltage, frequency, temperature, and attitude conditions. For optimum performance, however, the frequency should be kept within ten per cent of its rated value, and volts per cycle within a similar range.

As fast as instrumentation and control of aircraft in azimuth have posed new problems to solve, new techniques have been developed to meet the requirements. Magnetic compasses, stabilized compasses, and compass-controlled gyroscopes have all been steps in the evolution of direction indication but have failed in the past to be adequate for the violent maneuvers which modern fast airplanes can perform. It has been shown, however, that one form of compass-controlled directional gyroscope when supplied with the new "nontumbling feature" can provide satisfactory operation for these applications. Not only can this be accomplished, but the weight and size are such that the gyroscope indicator can be mounted in a standard $3^1/_8$-inch-diameter *AN*-size instrument panel cutout, a feature which is extremely desirable.

Appendix. The Servomechanism System of an Improved Azimuth Indicator

With reference to Figure 2, the primary detector of the directional measuring system is a permanent magnet which seeks a north-south direction in the remote compass transmitter *A*. A voltage pattern is produced which is transmitted to the compass repeater *E* and, if desired, can be transmitted to the directional repeaters *I*. The main function of the voltage pattern, however, is to act as the "desired behavior" of a regulatory loop or closed casual servo system. A paper on this subject, entitled "Unified Symbolism for Regulatory Control" by G. A. Philbrick, was presented at the December 1945 meeting of the American Society of Mechanical Engineers. Figure 11 shows a graphical or symbolic diagram of the servo system.

This voltage pattern produces an a-c flux in the Selsyn detector *B* that has a direction determined by the pattern. Using the symbolism of Brown and Hall[8] this direction can be represented functionally by $\Theta_i(t)$. The gyroscope itself positions pickup vanes in the Selsyn detector, the direction of which can be represented functionally by $\Theta_o(t)$. The difference between the compass reading (corresponding to the flux direction) and the gyroscopic reading (corresponding to the vane direction) then is represented by $\Theta_i(t) - \Theta_o(t)$ and this corresponds to the error signal voltage $\xi(t)$ induced in the pickup coil.

This error signal voltage $\xi(t)$ isthe input to a stabilizer control *C*, the outputs of which are direct currents. These currents flow through the coils of a torque motor *D* to produce a torque which can be represented functionally by $T_c(t)$. If the combined operator is $C_c(p)$, then the torque $T_c(t)$ is the result of $C_c(p)$ operating on the input $\xi(t)$. This output torque $T_c(t)$ combines with disturbance torques $T_o(t)$ such as those causing drifts to cause precession of the gyroscope in a direction to reduce the error signal. This precession operator $H_o(p)$ operating on the torques determines the position of the pickup vanes.

$$H_o(p)[T_c(t)+T_o(t)]=\Theta_o(t)$$

Figure 11 easily can be recognized as a standard simple servomechanism. The operator $C_c(p)$ can be considered a constant K for small values of $\xi(t)$ because in this region the stabilizer is a proportional amplifier, and the torque from the precession coils is essentially proportional to the current. However, for large values of $\xi(t)$, the amplifier saturates, and $C_c(p)$ has the necessary value to keep $T_c(t)$ or $C_c(p)\xi(t)$ essentially constant. The differential operator $H_o(p)$ is essentially an integral operator $1/p$ since the speed of gyroscope precession, or $(d\Theta_o)/dt$, is proportional to the torque $T_c(t) + T_o(t)$.

From this discussion it is apparent that the standard simple servomechanism equations which are covered fully in another paper[8] can be used to predict errors and stability of such systems.

The three basic equations can be written as follows, omitting (t) where convenient:

$$\left.\begin{aligned}&\xi(t)=\Theta_i(t)-\Theta_o(t)\\&\text{or}\\&\xi=\Theta_i-\Theta_o\end{aligned}\right\}\quad(1)$$

$$\left.\begin{aligned}&T_c(t)=C_c(p)\xi(t)\\&\text{or}\\&T_c=K\xi\end{aligned}\right\}\quad(2)$$

$$\left.\begin{aligned}&\Theta_o(t)=H_o(p)[T_c(t)+T_o(t)]\\&\text{or}\\&\Theta_o=1/p[T_c+T_o]\end{aligned}\right\}\quad(3)$$

Subtracting Θ_i from each side of equation 3 and substituting equation 2 in equation 3,

$$\Theta_i-\Theta_o=\Theta_i-1/p[K\xi+T_o]$$

which from equation 1 gives

$$\xi=\Theta_i-1/p\ K\xi-1/p\ T_o$$

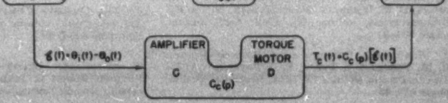


Figure 11. Graphical and symbolic diagram of compass controlled directional gyroscopic servo mechanism

or

$$\left.\begin{aligned}\xi &= \frac{\theta_i - 1/p\, T_o}{1+1/p\, K}\\ &= \left[\frac{p\theta_i - T_o}{p+K}\right]\end{aligned}\right\} \qquad (4)$$

The use of this denominator as an over-all system operator has been recognized in servomechanism design for years.

Thus $(p+K)^{-1}$ summarizes the system behavior over the range that K is a constant. Since $(p+K)^{-1}$ is the Laplacian transform for ϵ^{-Kt} its characteristic is an overdamped system without oscillation and with a time constant determined by K. From equations 1, 2, and 3

$$\theta_o(t) = \frac{C_o(p)H_o(p)\theta_i(t) + H_o(p)T_o(t)}{1+C_o(p)H_o(p)}$$

or

$$\theta_o = \frac{K\theta_i + T_o}{p+K}$$

or

$$\frac{d\theta_o}{dt} = K(\theta_i - \theta_o) + T_o$$

For "steady" conditions

$$\frac{d\theta_o}{dt} = 0$$

and

$$\theta_i - \theta_o = \xi = -T_o/K \qquad (5)$$

Equation 5 leads to the conclusion that the basic error of the system can be kept small by increasing the gain represented by K and by reducing the undesired torque T_o. Equation 4 which indicates a nonoscillating system with a time constant determined by K also indicates that K may be large without producing undesirable characteristics.

If K is to be large over the range that it is a constant, and if the precession rate is to be limited beyond this range in order to reduce errors that result from the gyroscope following the compass during the period that turn errors are present in the compass, a characteristic curve as shown in Figure 8 results.

With this characteristic, errors caused by normal undesirable torques are negligible and normal turn errors are small. If existing turn errors are objectionable, the compass control can be disconnected during turns.

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Some Aspects of Polyphase Motor Design—The Design and Properties of the Magnetic Circuit

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In a preceding paper,[1] an attempt was made to present a design point of view, indicating that the criterion for a design method is its direct ability to provide the construction data whereby desired machine specifications are met. The idea was illustrated by a procedure for polyphase motor design, stripped of refinements, and employing a magnetic circuit of fixed contour. Such a condition arises through the use of standard laminations in which designs are made. The questions naturally arise as to how these laminations are proportioned and what affect the proportions have on performance. It is with such matters that the present paper is concerned, directed chiefly at motors with semiclosed slots of 50 horsepower or less.

An investigation of the various proportions and dimensions which can be used in the magnetic circuit of an electric motor leads to some interesting relationships. Unfortunately, these readily can become of academic rather than practical interest unless several underlying facts of industrial and commercial importance are kept in mind.

Faced with the requirements of designing a suitable stator and rotor lamination for one definite horsepower and speed rating for a polyphase induction motor, it can be imagined readily that the relation of slot area to tooth area, and rotor diameter to stator diameter (and several other proportions) could be set up in useful ratios so as to obtain minimum copper, or minimum material cost, maximum efficiency, or any one of several other desideratum. This is a process of designing a lamination for a particular rating. In contrast to this is the process of designing a lamination for a given frame. The distinction is important enough to bear further explanation.

It has been pointed out that while a motor manufacturer sells horsepower, in reality he sets up to build frames. In some instances, a special frame or a special design may be made for one application, bearing one rating only, and planned for very large production. In such a case, the design of a "best" lamination for the case at hand may be justified.

In the majority of cases, a lamination is to be provided for a given frame and made as widely useful as possible for all ratings which expect to be sold in that frame size. Suppose, for instance, NEMA* frame 284 is considered. Standardized ratings in Table I indicate the horsepower and speeds which all are expected from this

Table I

Horsepower	RPM	Horsepower	RPM
General purpose, open type, 60 cycles			
10	3,600	1 1/2	600
7 1/2	1,800	1	514
5	1,200	3/4	514
3	900	3/4	450
2	720		
Totally enclosed			
3	1,800	2	900
General purpose, open type, 25 cycles			
5	1,500	2	500
3	750	1 1/2	500

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* National Electrical Manufacturers Association.

"container" of more or less fixed dimensions.

From a commercial point of view, all ratings shown should be obtained from exactly the same lamination, stacked to the same axial length. Such a practice would mean that any order for such ratings could be filled from a stock of built-up unwound bodies and fields, modifying only the stator windings and the end ring sections of the squirrel cage. This ideal state cannot be realized but it can be approximated, and the attempt has a great effect upon lamination design and the functions of the designer.

Considering the foregoing facts, an attempt to design a "best" lamination can be of interest only in the exceptional case. If the lamination is to be widely useful, the proportions determined as best for one rating readily may prove undesirable for another. The result is a series of compromises, based on the minimum number of laminations for any given frame, which will yield all of the standardized (and some special) ratings as required, without serious interference with the accepted standards of performance.

Dealing with lamination contours for such purposes, it is reasonable to take a pragmatic viewpoint. Accordingly, we will investigate first the proportions of a series of laminations which, through refinements of practice, have been known to yield useful results.

Method of Fixing Stator Proportions

A total of 108 stator and rotor laminations were used to determine lamination proportions. These varied in rotor diameter from 3 to 15 inches. Their proportions had been determined initially by investigations or as manifestations of the individual tastes of a number of designers over a period of years. As a first step in determining proportions, a series of points were plotted using diameter to the bottom of the stator slots versus rotor diameter. With surprisingly little variation these points fell on a straight line which can be expressed by the equation (see Figure 1)

$$D_1 = 1.175D + 0.647 \qquad (1)$$

To determine the outside diameter of the stator, relative flux densities of teeth and stator yoke will be used. Maximum tooth densities and average yoke densities will be determined.

Maximum stator tooth density is

$$B_{\max} = \frac{1.57 \times \text{poles} \times \text{flux per pole}}{\text{number of slots} \times t_w \times L} \qquad (2)$$

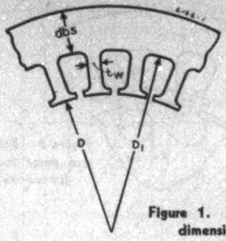

Figure 1. Stator dimensions

The tooth width (t_w) is assumed to be the narrow straight portion shown in Figure 1.

Average density in the yoke is

$$B_{\text{ave}} = \frac{0.5 \times \text{flux per pole}}{\text{radial yoke length} \times L} \qquad (3)$$

The radial yoke length or depth below slot will be abbreviated *dbs*.

As a working basis, it will be assumed that a reasonable magnetic circuit results from proportions giving a maximum stator tooth density of 120 per cent of the average yoke density. This ratio can vary somewhat with few bad effects. It is obvious that if tooth densities were kept constant on say both 4- and 8-pole designs, the yoke densities would reduce to half in the latter case. As will be shown in dealing with compromise laminations, over-all densities may well be adjusted. The effect of other ratios is investigated briefly in the appendix. From the 120-per-cent ratio,

$$\frac{1.57\phi P}{S_1 t_w L} = 1.20\frac{0.5\phi}{L(dbs)} \qquad (4)$$

Solving for depth below slot,

$$dbs = 0.382\frac{S_1 t_w}{P} \qquad (5)$$

Twice the depth below slot, plus D_1 represents the outside diameter of the stator laminations. It would be convenient to obtain this in terms of rotor diameter, and to do so the following method was employed.

The portion of the stator lamination lying just within the circle swept out by D_1 obviously fulfills the double purpose of supplying both the magnetic and the electric circuits. The division of this ring between slots and teeth affects the performance and results in an iron or a copper design. The effect is illustrated briefly in the appendix, but at this point it is sufficient to point out that a motor requiring low current densities (as a totally enclosed design with a cooling problem) might require a greater portion of this area for copper. Conversely, a design requiring high maximum torque, and intended for intermittent service, can be shown to require an increased tooth width. Our interest, however, is to determine a working value which is useful for general purpose designs, and hence the average for a group of existing laminations were investigated. When ($S_1 \times t_w$) versus rotor diameters were plotted for the 108 laminations, the results spread over a band which gave a clearly defined straight line. The equation is

$$S_1 t_w = 1.35D \qquad (6)$$

Substituting in equation 5,

$$dbs = 0.515\frac{D}{P} \qquad (7)$$

Then

$$OD = 1.175D + 1.03\frac{D}{P} + 0.647 \qquad (8)$$

Solving for D,

$$D = \frac{OD - 0.647}{1.175 + \dfrac{1.03}{P}} \qquad (9)$$

This equation, along with equation 1, is plotted in Figure 2 against rotor diameter. Based entirely upon average relationships observed from a number of laminations used in practice, and a few fundamentals, it fixes the main proportions of a stator lamination. Its validity as a working method will be checked later.

Compromise Dimensions

To illustrate the use of these curves or formulas, assume that laminations were to be provided for a frame which could accommodate a lamination diameter of ten inches. By these rules, Table II can be made.

Table II indicates that for each number of poles a different lamination would be necessary, and hence this practice represents the very requirement which is least desirable commercially. The point is, however, that by an examination of the dimensions, progressing in a logical manner, an intelligent compromise of dimensions can be made. Any such compromise

Table II

Poles	D, In.	D_1, In.	dbs, In.
2	5.58	7.14	1.42
4	6.53	8.32	0.84
6	6.94	8.80	0.595
8	7.16	9.06	0.46
10	7.32	9.25	0.377
12	7.42	9.36	0.317
14	7.49	9.45	0.276
16	7.55	9.52	0.242
18	7.59	9.56	0.217
20	7.62	9.60	0.196

involves a grouping of numbers of poles, and a consideration of mechanical strength of the yoke, in cases of many poles.

A 2-pole lamination usually represents an isolated design, not used for any other speed. This distinction is obvious, as the flux density in the yoke is only half of that obtained for a 4-pole machine, if the two designs are attempted in the same lamination. This change is always proportionately less as the poles increase in number. Table II illustrates this point.

No absolute rules can be laid down as to the grouping to be followed, partially because the relative production in various lines represents an individual problem. Thus compromise dimensions might be set up between the 4- and 6-pole designs, planning all production of a greater number of poles in one lamination. However, it might prove feasible to plan 4-, 6-, and 8-pole production in a compromise lamination, using a different design from 10-pole up. If 36 stator slots are used, fractional slot winding obviously would be required on the 8-pole designs. Grouping the 8-pole requirements with the lamination designed for the greatest number of poles, and using 72 slots, results in comparatively few fractional slot windings from 8 to 20 poles. Three-phase cases are emphasized here.

Granted that the grouping is an individual problem, it is of interest to observe the general effects on performance which would result in using a lamination designed by these rules for 4-pole use, when applied to 6-pole motors and vice-versa. Such results are tabulated in the appendix and are contrasted with other designs made in compromise laminations.

Stator Tooth Shape

Shape of the tooth tip and contour of the slot bottom are necessary for the completion of the stator lamination design. Relative values of d' and e' are important to the performance (see Figure 3). Too small a value of e' may make insertion of the windings difficult or impossible; too large a value increases the tooth flux fringing, the effective gap length, and the surface losses. An increased value of d' shows up chiefly in the increased slot leakage reactance.

A suggested working basis is to calculate

$$e' = 0.0143D + 0.0643 \qquad (10)$$

In designs with an unusually great number of slots for comparatively small diameters, the value of e' given by this formula may prove quite large. Any modification should be made towards reducing this value.

The depth d' should be chosen so that

$$d'/e' = 0.3 \text{ (approximately)} \qquad (11)$$

The bottom of the slot can be made semicircular, tangent to the circle swept out by D_1; or it can be made flat, with rounded corners. No recommendations can be made on this choice, aside from the comment that a die maker usually prefers the former practice.

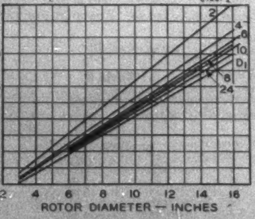

Figure 2. Values of outside diameter and D_1

The former varies with the number of poles when plotted as a function of rotor diameter

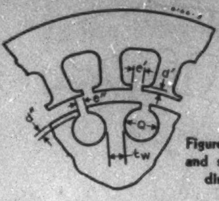

Figure 3. Rotor and stator tooth dimensions

Area of Stator Slots

The toroid between D and D_1 represents an area from which total tooth area can be subtracted to give gross slot area. Such a calculation gives the following expression for gross area of all slots:

$$\text{Area (sq in)} = 0.18D^2 + 0.758D + 0.33 \qquad (12)$$

Note that this is independent of the number of slots. Actually, because of rounding off corners, the net useful area per slot for laminations designed by these rules is approximately

$$\frac{0.9 \text{ gross area}}{S_1} = \text{actual slot area} \qquad (13)$$

Gross area is plotted in Figure 4.

Rotor Lamination Design

Items to be considered in the design of squirrel cage rotor laminations are the number and size of slots, and the dimensions of the slot openings. Common practice makes use of the following two forms of construction:

1. The round slot used with copper bars.
2. The deeper slot used with die cast cages.

The general problem will be considered in more detail for the first case.

It already has been pointed out that the stator slot opening can be obtained through the empirical relationship of equation 10. A useful value for the rotor opening is 70 per cent of e' or

$$e'' = 0.01D + 0.045 \qquad (14)$$

Modification usually should occur on the low, rather than the high side.

The slot constant K_s'' expresses the effective linkages per ampere per inch of length for a given slot shape. As a working basis, it will be assumed that this should always be 1.3. Now, since

$$K_s'' = 0.623 + d''/e'' \text{ (for round bars)} \qquad (15)$$

it follows that

$$d''/e'' = \frac{d''}{0.01D + 0.045} \text{ or } 0.677$$

Hence

$$d'' = 0.00677D + 0.0304 \qquad (16)$$

Both d'' and e'' are now fixed by these equations

The diameter of the slot remains to be determined. A working basis for deciding on slot diameter, and hence tooth width, involves a consideration of flux densities in the teeth. For field laminations the tooth width was fixed thus:

$$t_w = \frac{1.35D}{S_1}$$

In the same manner, if equal flux densities in the stator and rotor teeth are desired,

$$t_w'' = \frac{1.35D}{S_2} \qquad (17)$$

where S_2 = the number of rotor teeth or slots.

When round slots are used, the assumption that t_w'' indicates the tooth width at the narrow neck usually permits higher densities in the rotor than in the stator (see Figure 3). A figure of 118 per cent is useful and hence

$$t_w'' = \frac{1.35D}{1.18S_2} \qquad (18)$$

For cast rotors in which the tooth is commonly longer, equal flux densities in stator and rotor teeth are more desirable.

The magnetizing ampere turns required for rotor teeth otherwise may become excessive.

While the foregoing relationship enables the designer to determine the tooth width, a more direct calculation for fixing the diameter of the slot (and hence the bars) is very useful. This problem will be instigated further.

Our first concern is with an expression for the slot or bar circle diameter. A distinction must be made concerning diameter D as to whether it is used as the true rotor diameter or the stator bore. If considered as the stator bore, the air gaps immediately enter the calculations. While practices as to suitable gap lengths vary considerably, a reasonable basis is

$$\Delta = 0.0016 \times D + 0.0072 + 0.001L \qquad (19)$$

If modified, this usually would be increased for higher speed motors.

This gives the radial length of air gap in inches, as a function of both diameter and stack length. For our purposes it will be sufficiently accurate to assume a gap resulting from an average stack of three inches, leaving gap length as a function of diameter only.

To determine the slot diameter, we can make use of the following relationship:

$$\text{Diameter across slot centers} = D - 2\Delta - 2d'' - Q \qquad (20)$$

where Q = diameter of slot in inches. (This equation is in error to the extent that d'' is not measured on the diameter D, but is offset by the amount of one-half e''.)

Expanding equation 20, by equations 19 and 16 we obtain

$$\text{Diameter across slot centers} = 0.9833D - 0.0752 - Q$$

From which it follows that the diameter of the slot is

$$\frac{1.95D - 0.236}{S_2 + \pi} = Q \qquad (21)$$

It is of interest to note how the total area of the rotor slots varies as a function of their number. These areas are shown in Figure 5.

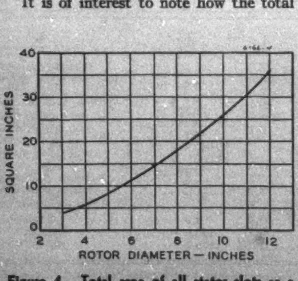

Figure 4. Total area of all stator slots as a function of rotor diameter

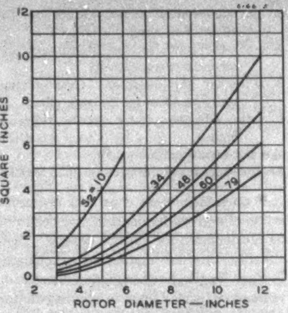

Figure 5. Total area of all rotor slots for various numbers of slots

Recapitulation

The foregoing rules are all that are required to determine the dimensions for stator and rotor laminations, once the outside diameter and the number of poles are known. Yet, it must be kept in mind that perhaps the greatest use of such formulas occurs through the design of general purpose or average laminations which are intended for use for a variety of designs, involving various numbers of poles. The assumption is that if one wishes to design a lamination for 4-, 6-, and 8-pole use the procedure would involve the preliminary design for 4- and 8-pole, and thus make an intelligent compromise between the known limits.

No data on recommended numbers of slots are given here as that involves other factors not relevant to this paper.

Lamination Capacity

It cannot be emphasized too strongly that once a set of laminations has been proportioned the maximum torques available from all motors built with these laminations have been fixed. Aside from the slight variation possible by a change in pitch, or the working of the magnetic circuit at various densities, a given stack of laminations results in a fixed value of maximum torque.

These facts give rise to the interesting possibilities of determining the inherent maximum torque of a given lamination when designed by the method evolved here and when designed according to any or no method.

It is possible to set up expressions for all motor constants in terms of the dimensions resulting from the rules given here. Then, by selecting a flux density for the teeth, the maximum torque for an entire line of laminations can be determined. Such calculations, while not shown here, can be used as the basis for empirical equations for maximum torque as functions of stack length and rotor diameter. By such a method, 4-pole 60-cycle designs of full pitch windings, and with a maximum stator tooth density of 96,000 lines per square inch, display maximum torques of

$$T_m = KL^{1.4}D^{1.87} \qquad (22)$$

where stator and rotor slots are, respectively, 24 and 34, K equals 0.21; 36 and 48, K equals 0.31; and 48 and 60, K equals 0.40.

Similarly, for 2-pole motors designed in these laminations according to the foregoing standards,

$$T_m = KL^{1.89}D^{1.68} \qquad (23)$$

where stator and rotor slots are, respectively, 10 and 12, K equals 0.10; 24 and 34, K equals 0.34.

These constants will vary as the square of the ratios of the densities, for any other values chosen.

It is of interest to compare the result of these empirical expressions with the actual design carried through in some detail in Appendix I. A maximum torque of 52.9 pound-feet was obtained therein. By equation 22

$$T_m = 0.31 \times 3^{1.4} \times 6.875^{1.87}$$
$$= 53.5 \text{ pound-feet}$$

Maximum Torque of Any Lamination

Calculation of the maximum torque of any polyphase induction motor involves the determination of the stator winding resistance, the leakage reactance, and the leakage factor. Basically, these all are calculated from certain physical dimensions of a lamination stack and the windings which can be provided in the laminations.

Such expanded expressions obviously can be substituted in the maximum torque equation to obtain an equation in terms of dimensions. As this substitution is straightforward, it requires no derivation here. The final equation is

$$T_m = 0.029\left(\frac{1 - P^2K_a}{K_d}\right)\frac{B_k^2L^2}{K_cP} \times \frac{1}{\dfrac{K_g}{f}\left(\dfrac{K_h}{P} + L + K_e\right) + K_aL + \dfrac{K_m}{P^2}} \qquad (24)$$

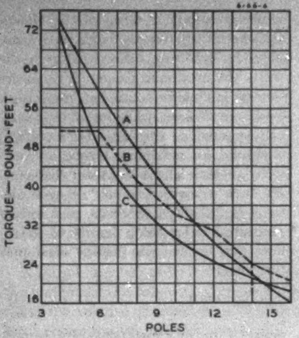

Figure 6

A—Inherent maximum torque of a lamination with constant flux density in the teeth

B—Maximum torque of standardized NEMA ratings (225 per cent ratio)

C—Inherent volt-amperes of a lamination with constant current densities. Scale not shown, as only the relative slopes and positions of the curves are significant

where

$$K_a=3.60(0.625p+0.375)\left(\frac{K_s'}{S_1}+\frac{K_s''}{S_2}\right)+\frac{7.5DK_z}{S_1S_2\Delta}(0.8p+0.2)$$

$$K_z=\left(\frac{t'+t''}{\lambda'+\lambda''}-0.5\right)^2$$

$$K_c=0.2D$$

$$K_d=0.845\frac{DK_w^2}{\Delta K_1}$$

$$K_1=\frac{\lambda'\times\lambda''}{t'\times t''}$$

$$K_e=\frac{328}{(K_w t_w S_1)^2}$$

$$K_g=\frac{0.153\times 10^6}{AkS_1}$$

$K_h=4.4p(D+\text{stator tooth length})$

$K_m=1.57D(1.25p-0.25)$

and

P = number of poles

p = pitch as a decimal fraction

B_k = maximum flux density in the stator tooth in kilolines per square inch

f = frequency in cycles per second

$K_s'+K_s''$ = primary and secondary slot constants, respectively

S_1+S_2 = number of primary and secondary slots, respectively

Δ = the radial air gap length in inches

$t'+t''$ = the effective tooth face in inches for primary and secondary, respectively. These are wider than the actual circumferential tooth widths by the effect of fringing

$\lambda'+\lambda''$ = the tooth pitches of primary and secondary, respectively

K_1 = the air gap constant. Greater than unity to allow for fringing effect

t_w' = the effective stator tooth width in inches

A = slot area in circular mils

k = the space factor for the slot (0.25 to 0.30)

Fortunately, these constants are properties of a given lamination and need not be recalculated for various designs. Specifically, they are unaffected by changes in stack length, frequency, flux density, or voltage. If the change in winding factor with the number of poles is neglected in making rough calculations for estimates, the constants are unaffected. To be more exact, K_a, K_d, K_e, K_h, and K_m must be adjusted for each change in pitch and winding factor.

A detailed example of this calculation is given in Appendix IV. However, its significance can be shown by assuming that one were interested in investigating design possibilities of everything from 4- to 16-pole motors in exactly the same lamination and stack. The inherent maximum torque of the lamination used in the appendix is shown plotted in Figure 6. The relation of such torques to the torque ratings required of the motors built in a corresponding NEMA frame also are indicated by these curves. Minimum ratios of maximum torque to full load torque are fixed at 200 per cent by NEMA, but the plotted values are at 225 per cent.

Volt-Ampere Rating of a Lamination

The volt-ampere capacity of a lamination is a fixed quantity only if allowable flux and current densities are chosen. The relationship can be derived as follows:

$$\text{Circular mils per conductor}=\frac{AkS_1}{nC} \qquad (25)$$

For a given maximum flux density in the teeth of B_k kilolines,

$$C=\frac{1}{L}\left(\frac{EP\times 70.9\times 10^3}{B_k f S_1 t_w K_w}\right) \qquad (26)$$

All terms have been defined previously.

If the allowable current density in circular mils per ampere is D_c, it follows that

$$\text{Volt-amperes (all phases)}=\frac{AkS_1^2 LB_k t_w f K_w}{D_c P\times 70.9\times 10^3} \qquad (27)$$

The foregoing formula holds for any lamination, but if the given rules are followed, the following simplifications result for laminations designed accordingly. Since

$$t_w=\frac{1.35D}{S_1}$$

$$\text{Volt-amperes}=DL\frac{AkS_1B_kfK_w}{D_cP}\times 1.9\times 10^{-5} \qquad (28)$$

Assuming an average space factor for the slot of 0.28, useful slot area can be expressed by the approximate relationship

$$AkS_1=0.28\times 0.9\times D^{1.41}\times 1{,}273{,}000$$

Then

$$\text{Volt-amperes}=D^{2.41}L\frac{B_k}{D_c}\text{rpm}\times K_w\times 0.051 \qquad (29)$$

Comparing this equation with the empirical expression for maximum torque, it becomes clear that as the diameter of the rotor (or correspondingly, the outside diameter of the laminations) increases,

Table III. Four-Pole Design

Lamination Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches
Full Load Torque: 7½-Horsepower 4-Pole Motor, 22.9 Pound-Feet

	1	2	3
Nominal rotor, D	6.875 in.	7.75 in.	6.5 in.
Depth below slot, dbs	0.885 in.	0.84 in.	1.0 in.
Tool width, t_w	0.257 in.	0.30 in.	0.225 in.
Slot constant, K_s'	1.295 in.	0.857 in.	1.3 in.
Slot area (sq in.)	0.355 in.	0.184 in.	0.400 in.
Wire weight (lbs)	15.6	8.1	17.6
Stator tooth density	96,000	96,000	96,000
Stator yoke density	80,000	98,000	62,000
Stator winding resistance	0.452 in.	0.64 in.	0.525 in.
Rotor slot constant, K_s''	1.3	1.3	1.3
Maximum torque	52.9	57.0	42.6
No-load values			
Amperes	6.3	8.0	5.04
Copper loss	53.6	123.0	40.0
Iron loss	150.8	178.0	128.7

Notes: Column 1—"Average" method.
Column 2—Large D, t_w.
Column 3—Small D, t_w.

the volt-ampere capacity of the laminations increases more rapidly than the maximum torque. The inverse of this has been noted by any designer of small polyphase motor parts, which are generally capable of yielding comparatively high maximum torque values. Attempting to give such parts their corresponding horsepower ratings can be done only at a sacrifice of circular mils per ampere.

The motors used in Appendix IV are calculated for inherent volt-ampere capacity and plotted in Figure 6. An allowance of 425 circular mils per ampere was used for the stator winding.

Appendix I

To illustrate the application of the formulas for the average lamination, assume that a lamination is to be designed for an outside diameter of 10.5 inches. As a working basis, the maximum stator tooth density of 120 per cent of average yoke density will be used.

From equation 7

$D = 6.875$ inches (4 poles)

Selecting 36 stator slots, we obtain the tooth width from equation 5:

$t_w' = 0.257$ inch
$dbs = 0.885$ inch

From equation 10, we obtain the slot opening:

$e' = 0.162$ inch

For $d/e = 0.3$,

$e = 0.049$ inch

"Rounding off" the corners of the slots yields a slot area of 0.355 square inch. This completes the stator lamination design, determined not as a fair lamination for say both 4- and 6-pole designs, but based on 4-pole calculations only.

For the rotor, using 48 slots and a gap of 0.025 inch (which is subtracted from the nominal diameter), the diameter of the rotor slot is 0.255 inch.

For semiopen slots, e'' equals seven-tenths of e', or 0.11 inch; and d'', when the rotor is ground, will be such that the slot constant is 1.3. The tooth width at the narrow portion is now 0.164 inch.

These laminations will be used for a design of a $7^1/_2$-horsepower motor, using the principles derived in the first paper on this subject.[1] Selecting a maximum to full load torque ratio of 225 per cent, and a suitable stator tooth density as 96,000 lines per square inch, the required stacking will be 3.0 inches.

For a full pitch winding, the complete performance is given in Table III, column 1.

The solution of the foregoing problem represents purely a routine task, of little significance beyond the indication that the lamination design method results in reasonable values. It leaves unanswered the following questions:

1. What would happen had a larger or a smaller rotor diameter been chosen?

2. What would be the effect of upsetting the ratio set up between tooth and slot areas in the stator?

To investigate, assume that a rotor diameter of 7.75 inches were chosen, along with a tooth width of 0.30 inch. The flux densities in stator core and teeth are now practically equal and are kept at 96,000 lines per square inch, as before. The lamination design, in general, is what would follow if 1.2 were changed to 1.0 in equation 4. The completed performance is tabulated in column 2 of Table III and bears out the conclusion that such a procedure results in a "flux motor" with high maximum torque, but overworked copper. Such a lamination design may represent a cooling problem, but might be useful for intermittent duty service with high torque demand.

In the other direction, suppose a lamination had been designed for a rotor diameter of 6.5 inches with a tooth width of 0.225 inch. The attempt to limit the flux density in the stator teeth to 96,000 lines now results in a weak motor, with maximum torque below the 200 per cent ratio if a $7^1/_2$ horsepower rating is expected. The increased slot area naturally results in less copper loss. Performance items are shown in column 3 of Table III.

Table IV. Six-Pole Design

Laminations Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches
Full Load Torque: 5-Horsepower 6-Pole Motor, 22.9 Pound-Feet

	1	2	3
Nominal rotor, D	7.315 in.	6.875 in.	7.315 in.
Depth below slot, dbs	0.628 in.	0.885 in.	0.628 in.
Tooth width, t_w	0.274 in.	0.257 in.	0.274 in.
Slot constant, K_s'	1.28	1.295	1.28
Slot area (sq in.)	0.386	0.355	0.386
Wire weight (lbs)	14.3	12.6	14.3
Stator tooth density	96,000	96,000	98,000
Stator yoke density	80,000	53,300	81,700
Stator winding resistance	0.659	0.814	0.630
Rotor slot constant	1.3	1.3	1.3
Maximum torque	50.2	42.9	52.6
No load values:			
Amperes	6.42	5.83	6.68
Copper loss	72.0	85.2	84.0
Iron loss	130.0	111.8	134.0

Column 1—"Average" method for 6-pole lamination.
Column 2—"Average" 4-pole lamination with 6-pole design.
Column 3—6-pole lamination with redesign of windings for higher value of maximum torque.

Appendix II

Examples given previously were based on the use of a lamination intended primarily for 4-pole operation. Using the same principles, but designing the lamination for 6-pole operation only, results in a rotor diameter of 7.315 inches with a tooth width of 0.274 inch. The resulting performance items are shown in Table IV, column 1. The 6-pole rating in this same stack represents a 5-horsepower motor. It will be noted that this motor is somewhat weaker in torque ratios than the one designed in Appendix I. A slight increase in densities brings about an equivalent value of maximum torque, with performance as shown in column 3.

While the data are not shown, a modification of this "average" lamination to yield equal densities in stator yoke and teeth, or to increase the tooth densities beyond the 120 per cent ratio setup, results in exactly the same trends already indicated in Appendix I.

In attempting to judge the validity of these lamination designs, one of the most effective means is to determine the effect of not following the rules. Thus, while average laminations have been designed for 4- and 6-pole use, it would be of interest to show how satisfactory a 6-pole design could be made in the lamination intended for 4-pole service.

The lamination of Table III, column 1 yields a 6-pole design as shown in Table IV, column 2. A flux density of 96,000 lines in the stator teeth again was chosen as a basis for calculation. As compared to the "average" lamination for 6-pole use, a sharp reduction in maximum torque results, along with an increase in copper loss.

This comparison bears out the fact that as an increased number of poles is used with a given lamination, the available torque always is reduced so long as tooth densities are maintained constant. This effect is exaggerated here, because the laminations had been deliberately designed for 4- or 6-pole duty respectively. In practice, the contrast would not have been so marked if the lamination dimensions had been compromised between 4- and 6-pole requirements, or if constant tooth densities were not maintained. Since an increase in the number of poles always results in reduced stator yoke densities (on a given lamination), an increase in tooth density then is not unreasonable. This may result in no great increase in core loss nor magnetizing current, and this practice does reduce the slope of the maximum torque versus number of poles curve for a given lamination.

Appendix III

It has been indicated in Appendix I that a 4-pole lamination for the outside diameter required should have a rotor diameter of 6.875 inches, and that the resulting performance was reasonable. Similarly in

Table V. Four- and 6-Pole Designs

Laminations Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches

	4-Pole		6-Pole	
	1	2	3	4
Full load torque	22.9 lb-ft	—	22.9 lb-ft	
Inside stator bore	7.00 in.	—	7.00 in.	
Depth below slot, *dbs*	0.826 in.	—	0.826 in.	
Tooth width	0.265 in.	—	0.265 in.	
Slot constant, K_s'	1.31	—	1.31	
Slot area	0.34	—	0.34	
Wire weight (lbs)	15.1	15.1	12.2	12.2
Stator tooth density	96,000	97,500	96,000	104,000
Stator yoke density	88,000	89,500	59,000	64,000
Stator winding resistance	0.374	0.385	0.708	0.602
Rotor slot constant, K_s''	1.3	1.3	1.3	1.3
Maximum torque	51.2	53.0	43.8	51.0
No load values:				
Amperes	6.6	6.8	6.1	7.3
Copper loss	51.0	55.0	83.0	101.0
Iron loss	155.0	161.0	116.0	129.0

Appendix II, a 6-pole lamination has been shown with a nominal diameter of 7.315 inches. Performance was acceptable, but when the 4-pole lamination was used on a 6-pole design, results were less satisfactory.

To illustrate that these rules are of aid in achieving a compromise design, let us assume that one lamination were to be used for both, and that its dimensions are selected as being between the two extremes.

Between nominal diameters of 6.875 and 7.315 inches, we will select 7 inches for the stator bore.

Between stator tooth widths of 0.257 and 0.274 inch, select 0.265 inch which follows from equation 6.

By equation 1, $D_1=8.847$ inches and $dbs=0.826$ inch.

The 4- and 6-pole designs made in this compromise lamination are tabulated in Table V. The significance of the statement that compromise laminations require adjustments in flux densities is apparent at once in examining columns 1 and 3 of this table. While the 4-pole motor suffers slightly in maximum torque, the 6-pole design is below the 200 per cent minimum, if the 96,000 density is used. Columns 2 and 4 show the results of adjusting the designs for the values of maximum torque obtained previously. The resulting performances do not compare too unfavorably with those obtained from the laminations made for individual designs.

Appendix IV

To illustrate the application of the inherent maximum torque of a lamination, we will consider a sample lamination designed without reference to the rules shown here, and displaying the following properties:

Outside diameter = 11.75 inches
Inside diameter = 7.50 inches
Slots = 48
Useful slot area = 123,000 cubic centimeters
Slot constant = 1.94
Tooth width = 0.23 inch
Tooth length = 1.125 inches
Rotor outside diameter = 7.445 inches
Air gap = 0.0275 inch
Slots = 60
Slot constant = 1.3
Gap constant = 1.27
B_k (assumed) = 85
Stack length, $L=3.00$ inches

Assume that this set of laminations was used for designs from 4 to 16 poles inclusive, using a fractional slot winding in the necessary cases. The initial calculations will be made with a pitch of 0.75 and a winding factor of 0.88 used as constant throughout.

Then

$K_a=0.271$
$K_c=1.50$
$K_g=2.60$
$K_m=8.10$
$K_d=141.0$
$K_e=3.45$
$K_h=28.5$

By equation 24,

T_m(4-pole design) = 73.8 pound-feet

Other values are plotted in Figure 6.

It is obvious that with the increase in the number of poles, the winding factor cannot remain constant, and the appropriate terms must be changed. The nature of the error resulting can be illustrated by comparing the results on the 16-pole design in which the winding factor is assumed first to be 0.88 and finally to be unity. In the latter case,

$K_a=0.3198$
$K_c=1.50$ (unchanged)
$K_d=180.0$
$K_e=2.72$
$K_g=2.60$ (unchanged)
$K_h=38.0$
$K_m=11.8$

T_m (16 poles) = 18.2 pound-feet versus 16 by uncorrected winding factor

The use of the same lamination for all of these poles hardly would be practicable because of the fractional slot windings which might result, and the poor slot combinations and unfavorable harmonic conditions on some designs. The curve of inherent maximum torque is chiefly useful to observe the trend. It must be kept in mind, also, that all designs were made for the same tooth density, and that with an increase in numbers of poles the yoke densities decrease. As previously mentioned it is sometimes practicable, and reasonable, to increase the densities slightly with pole increase. This would result in raising the lower portion of the curve.

As this set of laminations was intended for frame 284 it is of interest to compare the maximum torque expected of standardized ratings (4 to 16 poles) in this NEMA frame, with that inherent with this lamination. While standardization requires a maximum torque of only 200 per cent of full load torque, this curve is plotted with a reserve, using 225 per cent ratio. These are the points connected by dotted lines in Figure 6. Between this reserve, and the flattening effect of the inherent torque curve brought about by flux density adjustment with pole change, inherent maximum torques can be kept safely above the values expected of the standardized ratings. Similar investigations on other diameters of laminations indicates the theoretical soundness of the ratings called for in standardized frames. However, the inherent torque reserve of some horsepower ratings over others is very marked in a few cases.

Reference

1. Some Aspects of Electric-Motor Design—Polyphase-Induction-Motor Design to Meet Fixed Specifications, T. C. Lloyd AIEE Transactions, volume 63, 1944, January section, pages 14–20.

Effects of Thermal Characteristics of Aircraft Generators on Load Analysis

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Synopsis: The elaborate electric systems in modern long-range aircraft have made accuracy of load analysis more important than ever before. A more precise evaluation of the adequacy of proposed generating equipment can be made by averaging expected loads over several different periods which are chosen on the basis of the generators' thermal characteristics. These averages are compared graphically with curves which show the maximum overloads generators can tolerate without damage.

AS AIRCRAFT have increased in size and complexity, so have their electric systems. The problem of establishing a precise relationship between system size and the anticipated electric loads has become more and more important.

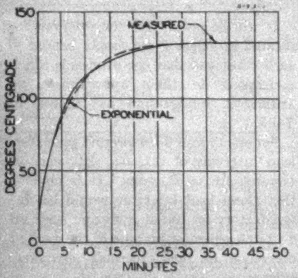

Figure 1. Simple exponential curve (broken line) compared with measured generator-winding heating curve (solid line)

Inadequate capacity may result in failure of the electric system with the consequent possible loss of the airplane, its crew, and its passengers. Too large a capacity, however, decreases pay load. Not only is extra fuel required to haul the useless weight, but additional fuel is needed to transport the fuel consumed in hauling this dead weight.

For airplanes with small electric systems, it may be sufficient to add up the power requirements of the various units of electric equipment and choose a generator which is estimated to be adequate. For somewhat larger systems, average power over various periods of time has been used as the basis for selecting generators. Later it was recognized by the more progressive manufacturers of large aircraft that peak loads and short time average loads are also important.

The most recent contributions to the making of an adequate load analysis include the following:

1. Loads should be averaged over not one or two, but several periods of time or "load intervals."
2. These periods of time depend on the overload thermal characteristics of the generator.
3. There is a maximum period of time over which loads may be averaged reasonably. The length of this maximum period depends on the generator.
4. A direct comparison between the maximum anticipated demand for electric power and the capacity of the generating system to meet that demand is essential to an adequate load analysis.

It is the purpose of this paper to explain why these propositions are true and to show how they are applied practically in determining the adequacy of an aircraft electric power system. The first step will be a consideration of the manner in which generator temperatures vary.

Temperature and Overloads

A typical aircraft generator is heated principally by the flow of electric current through its windings and is cooled by a blast of air under low pressure. The transfer of heat from the windings to the air follows the laws of thermodynamics, that is, the rise and fall of temperature occur exponentially. In most practical cases, the exact exponential function is complicated, but a simple function may be used as a first approximation. In Figure 1 a measured generator-winding heating curve (solid line) is compared with such a simple exponential curve (broken line). The close similarity between the two curves is apparent. In the analysis which follows it is assumed that the simple function is sufficiently accurate for practical purposes.

Experience has shown that a generator can stand a heavy overload for a short time without raising the temperature above a safe value. The less severe an overload is, the longer it can be tolerated without unwarranted damage. The length of time various overloads can be tolerated under a given set of operating conditions can be plotted. Figure 2 shows a typical curve. The curve is called a "permissible overload curve." The magnitude of overload (in terms of rated continuous-duty load) is plotted against the time for which the overload can be sustained without damage to the generator.

The permissible overload curve should be based on tests conducted by the manufacturer and should represent the guaranteed performance of the generator when subjected to overload. The curve rests on a specified set of conditions (generator speed, air pressure, air temperature, humidity, and so forth) and on assumed loading prior to the application of the overload. A family of such curves, representing performance under different conditions, would be even better. A conservative assumption is that thermal equilibrium under full-rated continuous-duty load has been attained before the overload is applied. Different conditions could be assumed, and curves based on them may be determined by test.

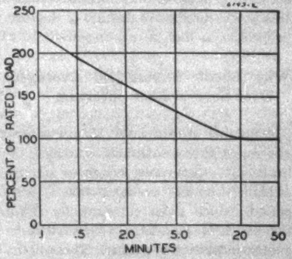

Figure 2. Typical permissible overload curve

Unfortunately, manufacturers generally do not supply permissible overload curves of the type here suggested. It therefore has been necessary to develop a method for estimating them as accu-

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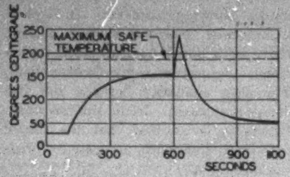
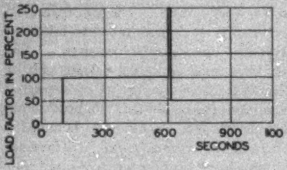
The approach utilized in the mathematical derivations in the appendix was suggested by Doctor George L. Shue, and the form and organization of the mathematical treatment is largely the work of B. D. Abramis, both of Consolidated Vultee Aircraft Corporation, San Diego, Calif.



A. Armature temperature curve



B. Corresponding load profile

Figure 3

The overload lasts five seconds. Load factor is the ratio of actual load to rated continuous-duty load

rately as possible. How this was done is set forth in detail in the appendix. Attention now is directed to the first of the four propositions which have been stated.

Why Loads Should Be Averaged Over Several Load Intervals

A number of load intervals are generally essential to a satisfactory analysis of loads on the generating system of a large airplane because averages taken over periods which differ substantially from the length of a damaging overload may give no indication of danger. This will be shown by means of two examples.

Figure 3A illustrates the variation of armature temperature with load. A cool generator is heated under rated load (as shown in Figure 3B) to full normal operating temperature. Thereupon a heavy overload is applied for five seconds, after which the load is reduced to one-half of full rated value. The attending rise and fall of temperature are shown in Figure 3A. It will be seen that during the overload the temperature quickly passes the maximum safe value and that about 25 seconds elapse before it decreases to a safe value. It will be noted that the graphs employ load factor as a parameter. It is the ratio of armature overload current (or power) to rated continuous-duty current (or power).

In Figure 4 the average load over intervals of 5, 15, 60, 180, and 600 seconds is plotted, the average always being taken so as to have the greatest possible value. The permissible overload curve is plotted to the same scale on this graph. The only average which exceeds the permissible overload curve is that for the 5-second interval. All the other averages lie on or below the curve and so do not indicate damage to the generator.

In Figures 5A, 5B, and 6, another case is illustrated in which a moderate overload is applied for 60 seconds. The armature reaches higher temperatures than in the first case and remains above the maximum safe value for more than a minute. This load is averaged over the same intervals of time as were used previously, and the averages again are compared with the permissible overload curve. Note that all the averages, except that for 60 seconds, lie below the permissible overload curve and give no warning of danger.

While the two cases chosen for purposes of illustration are not likely to be encountered in practice, experience has shown that when five well-chosen load intervals are used, excessive overloads may appear in only one or two of them, while the averages for the other load intervals appear to be safe.

From the foregoing considerations, it is evident that one or two periods of integration, regardless of their lengths, easily may fail to reveal the presence of damaging overloads. To discover with certainty every excessive overload, it would be necessary to use as many load intervals as there are different lengths of intermittent loads on the generating system; this would involve an impractical amount of work. Experience indicates that if the worst possible distribution of the intermittent loads is assumed, the use of four or five load intervals is usually adequate.

Load Intervals Chosen on the Basis of Permissible Overload Curve

If, several periods of integration must be used in order to make a load analysis which will reveal a dangerous overload, regardless of its duration, how long should each of the periods be? This question can be answered best by an example.

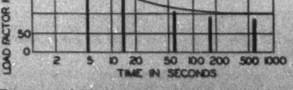


Figure 4. Values of load of Figure 3B averaged over periods of 5, 15, 60, 180, and 600 seconds

Figure 7 illustrates an overload characteristic curve for a large (40-kva) aircraft generator. Five periods over which loads may be averaged are indicated. These were chosen by the following criteria:

1. The shortest load interval should be somewhat longer than the shortest intermittent load worth considering and is usually of the order of five seconds. (Manufacturers sometimes supply overload ratings for such short times. It would be very helpful if a 5-second rating were furnished for all aircraft generators.)
2. The longest load interval should be equal to the shortest time for which the continuous-duty load ratings apply. Occasionally a manufacturer will supply this information. If he is unable to do so, the time may be estimated as being equal to three or four times the period of one cooling time constant.[*]
3. One of the intermediate load intervals should approximate the manufacturer's specified overload rating time (usually between one and five minutes).
4. The remaining intermediate load intervals should be so chosen as to make the differences between permissible overloads for successive intervals roughly equal.
5. The interval lengths chosen according to the foregoing rules should be modified enough to make the calculations easy.

It will be noted that the intervals indicated in Figure 7 are simply related to each other and that the overloads which correspond to them approximate an arithmetic progression as shown in Table I.

For one type of 6-kw aircraft generator, load intervals of 0.1, 0.5, 2.0, and 10 minutes are satisfactory. It is obvious that these load intervals would be unsatisfactory for use in analyzing loads on a 40-kva generator because of the marked difference in their thermal characteristics.

It follows, then, that before an adequate load analysis can be begun, the type of generator to be used first must be selected. Its thermal characteristics then are determined, and the load intervals are chosen. Loads may be averaged over each of these intervals, and the averages compared with the permissible overload (as indicated by the overload curve for the particular generator chosen). If the generators are not selected before the load intervals are chosen, a sufficient number of different intervals must be analyzed to insure adequate coverage of the overload characteristics of all types of generators which might be used.

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[*] This time constant may be determined experimentally as follows: The generator is heated under standard conditions until it reaches thermal equilibrium under full rated continuous-duty load. With the machine rotating and the standard conditions maintained, the load is removed, and armature temperature is plotted against time as the machine cools. The cooling time constant is approximately the time required for the armature temperature to drop through the first 63.2 per cent of the difference between cooling air temperature and the armature's full load operating temperature.

Maximum Time of Averaging Loads

Periods of one or two hours have been used in taking averages of loads on small generators. Some aircraft electrical engineers have been inclined to base all load intervals on the probable duration of each flight condition or aircraft operating condition (such as take-off, climb, cruise, land, and taxi) without regard to the thermal characteristics of the generators.

The duration of take-off, for example, is short and, for the analysis of loads during take-off, does justify limiting the longest load interval to a relatively short time. But it does not follow therefore that, because an airplane may cruise for hours, the longest load interval for use in analyzing cruising loads should be one or more hours. On the contrary, the maximum length of a useful load interval is dependent, as will be shown, on the thermal characteristics of the generating equipment.

Table I. Load Intervals and Corresponding Loads for 40-Kva Aircraft Generator

Load Interval Length, Min	Permissible Percentage of Full-Rated Continuous-Duty Load	Difference in Percentages
0.1	222	
0.5	190	32
3	158	32
15	125	33
60	100	25

As the generator's heat is transferred to the cooling air, its temperature drops unless new heat is added to offset the losses to the air. The transfer to the air of the heat present in the generator at any particular moment takes place exponentially. The amount of heat available for transfer is limited by the temperature difference between the generator and the air. In the simplest case, 63.2 per cent of this heat will be transferred to the air after one time constant, 86.5 per cent after two time constants, 95.0 per cent after three time constants, and 98.2 per cent after four time constants.** Thus the cooling of the generator causes it to "forget" how hot it was at an earlier time.

Whenever the load on the generator changes, the temperature rises or falls exponentially. If the new load is maintained for three or four times constants, generator temperature will reach a value substantially independent of its previous magnitude.

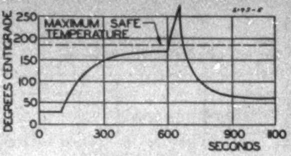
The fact that the generator "forgets" how hot or cold it was is illustrated in Figure 8, in which four different load and temperature conditions are graphed. In each case the abnormal load ends at the point marked zero time constants. The restoration of normal operating temperatures proceeds exponentially. After three time constants the four curves are very close together, and after five time constants they merge. These curves show that

1. Temperature at any particular time is not affected materially by the temperature at any time more than three or four time constants earlier.
2. The temperature at a particular instant depends on those loads and only those loads which occurred during the immediately preceding three or four time constants.
3. Averaging loads over a period greater than three or four time constants may result in hiding a damaging overload by averaging it with an underload so far removed in time that it could have no effect on temperatures reached during the overload.

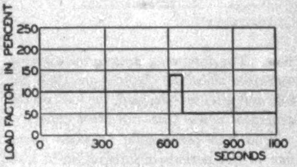
Choice of RMS or Average Values of Loads

A comparatively recent development in electric load analysis is the proposed use of rms values of load in place of average values. This proposal is based on excellent theoretical foundations. Since the load on aircraft generators is variable, its mean heating value is proportional to the average of the current squared. The suuare root of this value is a truer index of generator heating than is the average current. It has been found, however, that there is no significant difference between rms and average values in most practical cases. Since the rms values are much harder to calculate than average values, they rarely are justified in practice.

Direct Comparison of Anticipated Loads and System Capacity



A. Armature temperature curve



B. Corresponding load profile

Figure 5

Overload lasts 60 seconds

The permissibe overload curve indicates the magnitude of overload which, under specified conditions, may be applied for a given time without damaging the generator. A direct comparison of the curve with the greatest anticipated average (or rms) loads for each load interval is obtained by plotting the calculated averages to scale on the graph of the permissible overload curve.

Such a graphical comparison is illustrated in Figure 9. The average loads appear as vertical bars. The height of each bar is proportional to the average value of the load it represents. The location of the bar on the time axis of the graph indicates the length of the period over which the average was taken. One graph is made for each flight condition (take-off, cruise, and so forth) that is analyzed.

A glance at the graph indicates whether there are any dangerous overloads, and, if there are none, what margin of safety the chosen generator provides. By examining a set of graphs for the most important flight conditions, those conditions which merit special study may be chosen. These graphs convey a great deal of information in simple easily-understood form.

Load Analysis Procedure

After the type of generator to be used has been selected tentatively, a curve representing its thermal characteristics on overload is obtained or calculated. Such a curve is shown in Figure 9. This curve is based on the assumption that the generator is heated fully under full load before the overload is applied. The curve shows the time for which any given steady overload then may be applied without causing excessive damage to the insula-

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**In the expression $(1-\epsilon^{-t/B})$, where t represents time, B is the time constant. This expression states the portion of the transferrable heat which has been conveyed to the air in time t.

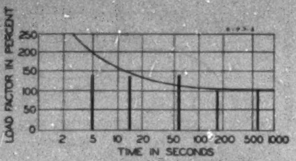

Figure 6. Values of load of Figure 5B averaged over periods of 5, 15, 60, 180, and 600 seconds

tion. The foregoing assumption is conservative since the chances are that, before and after an overload, the load on the generator is less than full rated capacity.

From the permissible overload curve four or five periods of integration or load intervals are chosen. These vary from a few seconds to 15 minutes or longer. They are selected in order to cover the range of permissible overloads in fairly even steps such as 100, 75, 50, and 25 per cent overload and no overload (full rated continuous-duty load). For convenience the lengths of the load intervals are made simple multiples or factors of each other.

The flight conditions or airplane operating conditions which are to be analyzed then are chosen. The most important ones are those in which the most severe electrical loads occur. These generally include take-off, cruise maximum, heavy combat, land, and ground operation.

Loads are tabulated and averaged for each flight condition and each load interval. The averages are plotted to scale as vertical bars on graphs of the overload characteristic of the generator. The resulting load analysis graph is illustrated in Figure 9. This graph not only summarizes the most important data, which conventionally are presented in tabular form on the load analysis chart, but also shows the relationship between calculated loads and available generator capacity.

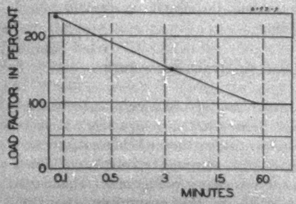

Figure 7. Permissible overload curve for 40-kva aircraft generator (calculated by method described in the appendix)

Five load intervals suitable for analyzing loads on this generator are indicated

While the propositions and method explained were applied to the analysis of loads on aircraft generators, the same principles may be used in analyzing many other types of electrical loads.

Appendix. Estimating the Permissible Overload Curve

Because manufacturers at present generally do not supply curves showing the time for which various overloads may be applied safely to the generator under specified conditions, a mathematical analysis of the thermal behavior of overloaded aircraft generators was undertaken.

The following simplifying assumptions have been made:

1. Heating and cooling occur as simple functions.
2. The armature heats and cools uniformly, and temperature gradients usually may be disregarded.
3. The highest temperatures occur in the armature winding.

An exact evaluation of the effect of field heating on armature temperature and of radiation and metallic conduction losses is not attempted.

The analysis made in this paper is necessarily approximate. The goal toward which it is directed is the expression of the time for which a given overload may be tolerated in terms of the load factor L (the ratio of actual load to rated continuous load) and a set of constants which can be evaluated. The analysis is divided into five operations.

1. Solution of the differential equation for rate of temperature change.
2. Expression of the temperature T in terms of the load factor L and time t.
3. Obtaining the expression for the permissible duration of a given overload.
4. Evaluating the ratio of the effective thermal capacity C to the coefficient of cooling K and incorporating the result into the final equation for permissible overload duration.
5. Drawing the permissible overload curve.

All temperatures are expressed in degrees absolute. The symbols and units used in the course of the analysis are explained as they occur.

Analysis of the thermal behavior of the armature winding is approached best by considering the rate of change of temperature. The truth of the following statement is evident:

$$\text{Rate of change of temperature} = \frac{\text{rate of heating} - \text{rate of cooling}}{\text{effective thermal capacity}}$$

Stated in mathematical terms, this becomes

$$\frac{dT}{dt}=\frac{HT-K(T-T_a)}{C} \tag{1}$$

where

T = Kelvin temperature of the armature winding
t = time in seconds
H = rate of heating per degree absolute and equals I^2R'/T' watts per degree Kelvin
I = armature current in amperes
R' = armature resistance in ohms at reference temperature T'
K = coefficient of cooling (watts per degree Kelvin)
T_a = Kelvin temperature of the cooling air
C = effective thermal capacity (joules per degree Kelvin)

This equation is based on the following reasoning:

1. For permissible generator temperatures the resistance R of copper varies roughly as its absolute temperature, $R=R'(T/T')$. Hence the rate of liberation of heat is proportional not only to the current squared, but also to the absolute temperature of the copper, $I^2R=(I^2R'/T')T$. The rate of heating is, therefore, HT.
2. The rate of cooling follows Newton's law of cooling and is proportional to the difference in temperature between the winding and the cooling air. It is represented by the expression $K(T-T_a)$ in which K is the factor of proportionality. This factor is presumed to be constant for a given installation under a given set of cooling conditions and is independent of H.
3. The rate of change of temperature varies inversely as the effective thermal capacity C.

Inasmuch as it is difficult to evaluate the constant K in practical cases, a solution in terms of temperature and load is desirable. The analysis will proceed as follows:

1. The differential equation 1 is solved (equations 2 and 3).
2. In the following steps (through equation 11), the ratio H/K is expressed in terms of the load and known temperature constants.
3. After expressing the temperature equation in a simplified parametric form (equation 13), the equation is solved for time t' (equation 14). The new equation contains a factor C/K which remains to be evaluated. This is accomplished by substituting two known boundary conditions in the overload time expression and solving the resultant equations simultaneously.

Operation I—Solution of the Differential Equation

$$\frac{dT}{HT-KT+KT_a}=\frac{dt}{C} \tag{2}$$

The solution of this equation is

$$T=\left(T_o+\frac{KT_a}{H-K}\right)\epsilon^{(H-K)t/C}-\frac{KT_a}{H-K} \tag{3}$$

where T_o is the initial temperature.

Note, that when $H=K$, equation 2 becomes

$$\frac{dT}{KT_a}=\frac{dt}{C} \tag{2a}$$

the solution of which is

$$t=\frac{C}{K}\times\frac{T-T_o}{T_a} \tag{3a}$$

Operation II—Expression of T in Terms of L and t

In the following steps equations 3 is expressed in terms of the ratio H/K. This ratio then is stated in terms of the load factor L and of S, a function of known temperature constants. The variable H thus is eliminated.

Letting $P=H/K$, equation 3 becomes

$$T=\left(T_o+\frac{T_a}{P-1}\right)\epsilon^{K(P-1)t/C}-\frac{T_a}{P-1} \tag{4}$$

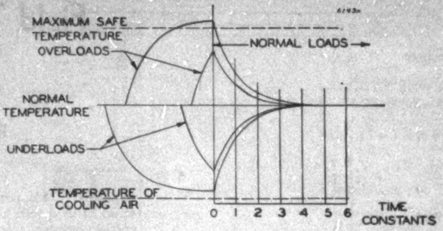

Figure 8. Temperature versus time curves showing that, regardless of deviations caused by underloads and overloads, temperatures return to normal a few time constants after normal load is restored (at time equals zero time constants)

When P is less than unity, that is, when H is less than K, the exponential term vanishes as t approaches infinity. This is clearly the case when the generator is carrying its full rated load. Letting the subscript n denote value at full rated continuous-duty load, the final steady state temperature is T_n

where

$$T_n = T_a/(1-S) \tag{5}$$

and

$$S = P_n = H_n/K = (T_n - T_a)/T_n \tag{6}$$

For overload conditions it may be assumed that H, the power consumed per degree absolute in heating the armature winding, is proportional to the square of the armature current. Letting L equal the ratio of load current to rated continuous-duty load current,

$$H = QL^2 \tag{7}$$

where Q is the factor of proportionality.

At full rated load, L equals 1, and H has its "normal" value of H_n. Therefore

$$Q = H_n \tag{8}$$

and

$$H = H_nL^2 \tag{9}$$

Substituting equation 6 in equation 9 and dividing each side by K,

$$H/K = H_nL^2/K = SL^2 = P \tag{10}$$

Substituting equation 10 in equation 4,

$$T = \left(T_0 + \frac{T_a}{SL^2-1}\right)\epsilon^{(K/C)(SL^2-1)t} - \frac{T_a}{SL^2-1} \tag{11}$$

Let T_s be the asymptote of this equation. Its value is that of the constant term of equation 11.

$$T_s = \frac{T_a}{1-SL^2} \tag{12}$$

Using the parameter T_s, equation 11 becomes

$$T = T_s - (T_s - T_0)\epsilon^{-(K/C)(T_a/T_s)t} \tag{13}$$

Note, that for the special case in which $H = K$,

$$T = T_0 + T_a(K/C)t \tag{13a}$$

Operation III—Solution for the Permissible Duration of an Overload

Solving equation 13 for time t when T equals T_p, the peak permissible winding temperature, and T_0 equals T_n, the normal full rated continuous load (assuming that the generator is fully heated under rated continuous load before the overload is applied),

$$t' = \frac{C/K}{T_a/T_s} \ln \frac{T_s - T_n}{T_s - T_p} \tag{14}$$

For the case in which $H = K$,

$$t' = (C/K)(T_p - T_n)/T_a \tag{14a}$$

where t' is the permissible overload duration, that is, the time the given overload will require to raise armature temperature from its normal value T_n to the peak permissible value T_p, and T_s is the variable (see equation 12).

When t' is plotted against the load factor L, it is seen that the curve is asymptotic to the line

$$L = \sqrt{(T_n/T_p)(T_p - T_a)/(T_n - T_a)}$$

Therefore equation 14 is useful only for load factors greater than this value. In practice typical usable values of L usually range from 1.5 to 2.5.

Operation IV—Evaluation of C/K and Statement of Final Solution

Attention now is directed to the quantity C/K. In the foregoing equations the thermal capacity C was assumed to be a constant. Actually, the effective thermal capacity is a function of both t and L. Taking into account the variation of C with time would have resulted in a more complicated differential equation and solution. Because of the uncertainty as to the nature of that variation, equations including this variation (some of which were set up and solved) were found to be of little value. However, treating C as a function of the load factor L only, does not affect the preceding analysis. The results obtained on this basis are in better agreement with experimental evidence than is an oversimplified solution in which the thermal capacity C is regarded as a constant.

The maximum value of C cannot exceed the thermal capacity of the material heated by the generator. There also must be a minimum value of C which holds only for a very brief period immediately following the application of a large overload. In so short a period there is no time for heat to be transferred from the copper of the armature winding to surrounding material. Because of variations in the resistance of different parts of the armature winding (a result of temperature differences for example) the effective minimum value of C may be far below the thermal capacity of the entire armature winding. The rate of heating at the hot spots is accelerated by the local increase in resistance caused by the increase in temperature.

Several assumptions as to the nature of the variation of C/K (K being a constant) were tried. The resultant permissible overload curves were plotted for a 40-kva aircraft generator for which two overload ratings were specified, one for five seconds and the other for five minutes. The following assumptions were evaluated from an examination of the curves:

1. C/K = constant.
2. $C/K = ML$.
3. $C/K = M/L + N$.
4. $C/K = M/(L + NL^2)$.
5. $\log C/K = N(M - L)$.

M and N represent constants, and L is the load factor. Only the last of these assumptions was found to be workable. Each of the other four either produced negative values of C/K for some range of values of L or failed to permit the permissible overload curve to pass through both specified rating points.

Letting t_q and t_r denote the durations of the two specified overloads and letting L_q and L_r respectively represent their load factors, by simultaneous equations we find

$$M = \frac{aL_r - bL_q}{a - b} \tag{15}$$

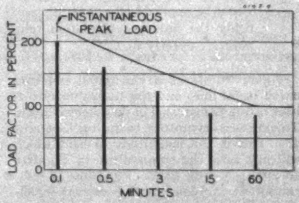

Figure 9. Typical load analysis graph

Bars represent average (or rms) values of load taken over periods of 0.1, 0.5, 3.0, 15, and 60 minutes

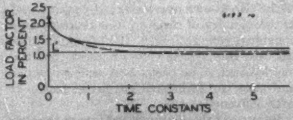

Figure 10. Illustration showing manner in which estimated (broken line) curve is joined to the calculated (solid line) permissible overload curve

The curve which is used consists of the estimated curve plus that portion of the calculated curve which lies to the left of the junction of the two curves

$$N = \frac{a-b}{L_r - L_q} \tag{16}$$

where

$a = \log (t_q/U_q)$
$b = \log (t_r/U_r)$

When $L = L_q$

$$U_q = \frac{T_s}{T_a} \ln \frac{T_s - T_n}{T_s - T_p} \tag{17}$$

When $L = L_r$

$$U_r = \frac{T_s}{T_a} \ln \frac{T_s - T_n}{T_s - T_p} \tag{18}$$

To evaluate equations 17 and 18, refer to equations 12 and 6.

Operation V—Drawing the Permissible Overload Curve

The permissible overload curve is calculated on the basis of known values for the following quantities:

T_a = absolute temperature of cooling air
T_n = normal hot spot insulation temperature (absolute) for rated continuous duty load
T_p = maximum permissible insulation temperature (absolute)
t_q = time for which an overload of load factor L_q can be applied to the generator and bring the hot spot insulation temperature exactly to T_p*
t_r = time for which an overload of load factor L_r can be applied to the generator and bring the hot spot insulation temperature exactly to T_p*

Figure 10 shows a calculated overload curve (solid line) and the two known overload rating points used in its determination. This curve is asymptotic to the line L-equals-L'. This is the magnitude of load which will just raise the temperature to T_p after an infinitely long time. Equation 14 indicates that this load can be carried indefinitely, but the generator manufacturer's rating for continuous duty is represented by the line L equals 100 per cent. For reasons already discussed, after three to four time constants have elapsed, the manufacturer's rating applies (refer to Figure 8). Hence the estimated curve (broken line) is added. It joins the calculated line below the lower of the two known overload rating points and in the vicinity of three to four thermal time constants merges with the L-equals-100-per-cent line. It is recommended that the overload curve be plotted on semi-log paper and that the estimated curve be drawn in with a French curve.

Because of the composite nature of the permissible overload time curve, an inflection point may occur in the vicinity of the joint between the estimated and calculated sections. This point is especially likely to happen when the curve is plotted on semi-log paper

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*In practice it may be assumed that t_q and t_r are the rated times for which overloads L_q and L_r may be applied.

Field Tests on Power-Line Carrier-Current Equipment

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Synopsis: This paper briefly describes a series of field tests made on the power line carrier current telephone system of the Pacific Gas and Electric Company, covering both amplitude and frequency modulated equipment. A discussion of various tests and test results is given, with a comparison of performance of the amplitude and frequency modulated equipment.

THE Pacific Gas and Electric Company has used carrier current telephone equipment on its high voltage transmission network for 23 years. Severe noise problems exist on some parts of the 230-kv transmission system because the lines have high corona losses. Early attempts to provide satisfactory carrier current communication over these lines were unsuccessful and led to the development of single-side-band power-line carrier-telephone equipment, some of which has been in service since 1927 and has been described previously.[1]

Major additions of hydroelectric generation and transmission facilities were made in 1944 requiring a considerable extension to the carrier current communication system. As some of the channels of the new carrier telephone system were to operate over the high corona lines, the noise problem required careful consideration. The use of frequency modulated carrier equipment[2] was proposed as a solution to the noise problem and an installation of this type equipment was made over a part of the system.[3] The complete carrier current telephone additions made in connection with the power system expansion in cluded four 2-frequency-duplex amplitude-modulated channels and one 2-frequency-duplex frequency-modulated channel on the Pacific Gas and Electric Company system. In addition, another 2-frequency-duplex amplitude-modulated channel was installed by the United States Bureau of Reclamation between Shasta Dam power house and the Pacific Gas and Electric Company Shasta substation. The carrier current telephone channels on the 230-kv system, with connecting channels, are shown in Figure 1.

The power system expansion was made to meet wartime requirements, and it was impossible to make complete tests of the carrier current equipment in the field at the time of installation because of the shortage of personnel, test equipment, and time. After the installation was completed, it was thought that maximum performance was not being realized from the system. Listening checks resulted in differences of opinion on the performance of various parts of the system. An extensive investigation of the characteristics of the system was undertaken to obtain quantitative data of the performance, and to make any adjustments found necessary to provide improvements.

Description of Tests

Data were taken to determine

1. The carrier frequency response of the equipment.
2. The audio frequency response.
3. The effects of transmitter output power level on the received signal and signal-to-noise ratio.
4. A comparison of frequency and amplitude modulation signal-to-noise ratios.
5. The effects of percentage modulation on signal-to-noise ratio.
6. Audio frequency attenuation of wire line and cable extensions used with the carrier current equipment.

The carrier frequency response gives an indication of the correctness of the tuning of the transmitter output stages, the line tuning units, and the receiving equipment. Carrier frequency response data were obtained by reading the transmitter output voltage, the coupling capacitor current, carrier output current (measured at the transmitter), received voltage, and receiver limiter current as the transmitter oscillator frequency was varied. From

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R. H. MILLER, an assistant engineer with the Pacific Gas and Electric Company, San Francisco, Calif. is now on a temporary assignment with the United States Navy.; E. S. PRUD'HOMME is an electronics engineer with the General Electric Company, San Francisco, Calif.

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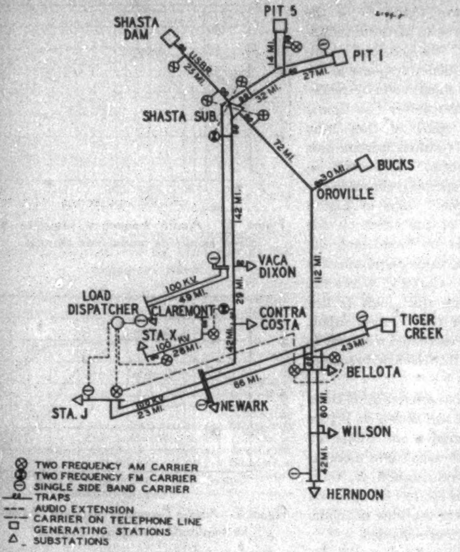
SHASTA DAM
PIT 5
PIT I
USBR
25 MI.
14 MI.
27 MI.
32 MI.
SHASTA SUB.
72 MI.
BUCKS
30 MI.
OROVILLE
142 MI.
112 MI.
VACA DIXON
100 KV
49 MI.
LOAD DISPATCHER
CLAREMONT
29 MI.
CONTRA COSTA
TIGER CREEK
STA. X
100 KV
26 MI.
42 MI.
43 MI.
BELLOTA
66 MI.
STA. J
100 KV
23 MI.
NEWARK
60 MI.
WILSON
42 MI.
HERNDON
⊗ TWO FREQUENCY AM CARRIER
⊗ TWO FREQUENCY FM CARRIER
⊖ SINGLE SIDE BAND CARRIER
TRAPS
AUDIO EXTENSION
CARRIER ON TELEPHONE LINE
□ GENERATING STATIONS
△ SUBSTATIONS

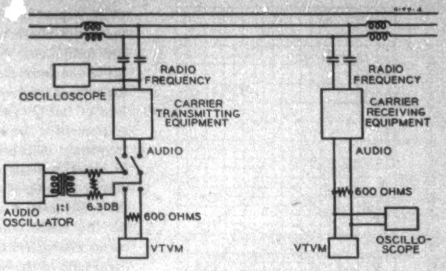
RADIO FREQUENCY
OSCILLOSCOPE
CARRIER TRANSMITTING EQUIPMENT
AUDIO
AUDIO OSCILLATOR
1:1
6.3DB
600 OHMS
VTVM
RADIO FREQUENCY
CARRIER RECEIVING EQUIPMENT
AUDIO
600 OHMS
OSCILLO-SCOPE
VTVM

Figure 2 (above). Circuit used to obtain audio frequency response characteristics

Figure 1 (left). Carrier current telephone channels on the Pacific Gas and Electric Company 230-kv transmission system

these data it was possible to plot and observe the frequency response of each part of the equipment through which the carrier frequency signal passed.

Audio frequency response was obtained by supplying constant level audio frequency tones to the transmitter modulator and measuring the receiver output voltage. These tests were carried out after the correct carrier frequency adjustments had been established. The circuit used for the audio frequency response measurements is shown in Figure 2. These measurements were made on each channel of the system and over several channels in cascade. The input level of the audio tones was set to provide a high percentage modulation, approximately 75 per cent for normal voice, as determined by observation with an oscilloscope of the carrier frequency output. With this setting the modulation limiter would prevent overmodulation on peaks. Both the carrier frequency and audio outputs were observed with an oscilloscope to determine whether or not distortion was present.

Noise measurements were made on all channels and the signal-to-noise ratios were determined. Noise and signal measurements were made with a vacuum-tube voltmeter, and noise measurements are unweighted. Weighted noise measurements were taken on some channels, but an insufficent number are available for comparison and will not be presented.

Received signal and signal-to-noise ratios were observed as the transmitter power output was varied.

Frequency and amplitude modulation signal-to-noise ratios were compared on the same power transmission line section in order to obtain actual field data on the effectiveness of frequency modulation as a means of overcoming noise in carrier current telephone systems.

Audio frequency attenuation of the audio frequency extensions on the carrier system were obtained. As the extensions were of considerable length, their characteristics were considered to be a factor in over-all performance.

Results

It was found that considerable improvement was possible in some parts of the system, and data and curves showing the characteristics of some of the channels, both before and after readjustments, will be presented.

CARRIER FREQUENCY RESPONSE

Figure 3 shows the carrier frequency response of the frequency modulation channel with its original adjustment. These curves show a considerable dissymmetry about the 75-kc operating frequency, which was apparent as distortion in the receiver output. Careful retuning of the line tuning equipment resulted in doubling the transmission line current, a considerable increase in received voltage, and symmetrical variations about the operating frequency as indicated by the curves of Figure 4.

AUDIO FREQUENCY RESPONSE

Considerable audio frequency response data were obtained. The manufacturer had guaranteed that the audio characteristics of the individual channels would be within plus or minus five decibels between 300 and 3,000 cycles. It was possible to come well within these limits on each channel. Audio characteristics of the 75-kc frequency modulation channel operating between Shasta and Contra Costa substations are shown in Figure 5. The audio response curve for the 85-kc amplitude modulation channel between Shasta substation and Pit 5 power house is illustrated in Figure 6. This curve is typical of the response of the amplitude modulation equipment.

One of the amplitude modulation re-

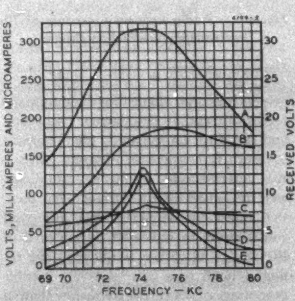

Figure 3. Carrier frequency response of 75-kc frequency modulation channel with improper line tuning

A—Output current at transmitter (milliamperes)
B—Transmitter output voltage
C—Coupling capacitor current (milliamperes)
D—Received voltage
E—Receiver limiter current (microamperes)

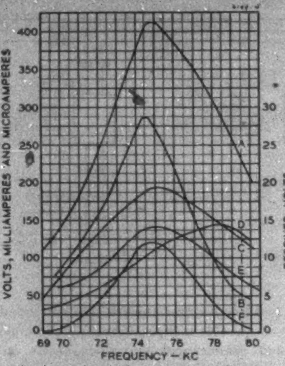

Figure 4. Carrier frequency response of 75-kc frequency modulation channel with correct line tuning

A—Output current at transmitter (milliamperes)
B—Received voltage (high *Q*)
C—Coupling capacitor current (milliamperes)
D—Transmitter output voltage
E—Received voltage (low *Q*)
F—Receiver limiter (high *Q*) current (microamperes)

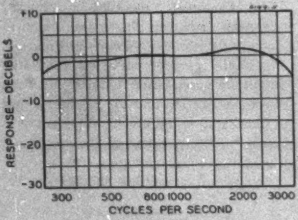

Figure 5. Audio frequency response of 75-kc frequency modulation channel

ceivers had been aligned for maximum detector current at the carrier frequency, and the effect on the audio characteristic is shown in curve *A* of Figure 7. When the intermediate frequency amplifiers were realigned for the correct band pass, the audio characteristic was improved considerably as indicated by curve *B* of Figure 7. At carrier frequencies the required band pass is a relatively greater percentage of the carrier frequency than in ordinary space radio applications. As in any highly selective receiver employing band pass circuits, tuning adjustments must be made carefully.

A limiting factor in audio frequency response, in this case of 2-frequency-duplex carrier system with interphase coupling using only 0.001-microfarad coupling capacitors, was found to be the resonance characteristics of the line tuning equipment. It was possible to improve the audio frequency response at the expense of received signal volts by reduction of the *Q* of the receiver line tuning equipment. Broadening of the audio frequency response resulted in some loss in signal-to-noise ratio. The effect on carrier frequency response is shown in the "low *Q* received volts" curve of Figure 4. An indication of the effect on the audio characteristics by broadening the band pass of the line tuning equipment is shown in Figure 8. Curve *A* shows the audio characteristics with high *Q* line tuning elements on the receiver input. Curve *B* shows the increased response to high frequencies with broadened receiver line tuning.

Over-all audio characteristics of three channels in cascade are shown in Figure 9. This system includes one frequency modulation and two amplitude modulation channels between station *X*, Oakland, and Pit 5 power house. The over-all response nearly meets the plus or minus 5-decibel limits set for one channel.

POWER OUTPUT

The effects of the variation of transmitter output on the received signal and signal-to-noise ratio for the 75-kc frequency modulation channel are given in Table I. Received volts in Table I are signal plus noise. These results indicate the effectiveness of the frequency modulation receiver in maintaining the signal-to-noise ratio with large variations in the transmitter output and the received signal. It was possible to maintain communication over a 175-mile 230-kv line when using only the exciter unit of the frequency modulation transmitter with a power output of two watts.

FREQUENCY AND AMPLITUDE MODULATION SIGNAL-TO-NOISE COMPARISON

A comparison of frequency and amplitude modulation signal-to-noise ratios is given in Table II. From these data it is apparent that the frequency modulation equipment is capable of maintaining a high signal-to-noise ratio even as the carrier signal approaches the noise level, and consequently communication can be maintained under conditions when a conventional amplitude modulation system would be inoperative.

Table I

Transmitter Output, Volts	Receiver Input, Volts	Signal-to-Noise Ratio, Decibels
50	3.6	29.5
25	2.2	27.8
10.8	1.6	24.5
4.8	1.5	24.5

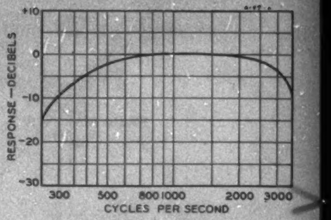

Figure 6. Audio frequency response of 85-kc amplitude modulation channel

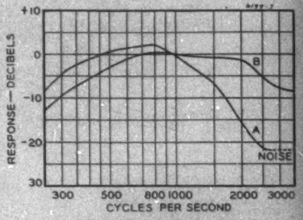

Figure 7. Audio frequency response of 140-kc amplitude modulation channel

Curve *A*—Receiver aligned for maximum detector current at carrier frequency
Curve *B*—Receiver aligned for correct band-pass

EFFECTS OF MODULATION LEVEL ON SIGNAL-TO-NOISE RATIO

On the frequency modulation channels with a deviation ratio of less than one-to-one, the signal-to-noise ratio was reduced considerably. Checks were made with deviation ratios of approximately one-third to one, and one-to-one, with the result that the signal-to-noise ratio was increased from 22 decibels with the lower to 34.5 decibels with the higher level of modulation. With a small deviation ratio, the phase modulation resulting from noise becomes appreciable relative to the desired frequency modulation and the noise output is proportionally higher. With the modulation correctly adjusted for normal voice input, overswing on peaks is limited by means of the modulation limiter which tends to maintain the deviation ratio near the correct value. A discussion of deviation ratios, band-width requirements, and noise considerations is given in reference 2. One of

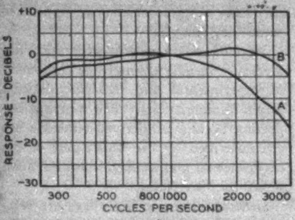

Figure 8. Audio frequency response curves showing the effect of the band pass of line tuning equipment

Curve *A*—High *Q* line tuning equipment
Curve *B*—Low *Q* line tuning equipment

Table II

Transmitter Output, Watts	Frequency Modulation Signal-to-Noise, Decibels	Amplitude Modulation Signal-to-Noise, Decibels	Frequency Modulation Over Amplitude Modulation Decibels
10	36.2	22.5	13.7
5	36	22.5	13.5
1	32.4	22.4	10.0
0.2	24.8	17.5	7.3

the early difficulties in the field adjustment of the frequency modulated carrier equipment was lack of a simple field check of deviation. With amplitude modulation equipment it is possible to observe the modulation envelope with an oscilloscope to check percentage modulation and the presence of distortion. A similar procedure for the frequency modulation equipment was suggested by R. W. Beckwith. With the carrier output connected to an oscilloscope with internal synchronization, a change of frequency will cause the pattern to move horizontally on the screen. By calibrating the horizontal movement versus frequency swing, a simple but sufficiently accurate method of measuring deviation is obtained, and the performance of the frequency modulation transmitter can be observed conveniently.

Audio Characteristics of Wire and Cable Extensions

The carrier terminals are located at high voltage substations at a considerable distance from the dispatcher and general offices of the company. The carrier telephone system was installed primarily to provide communication for the dispatchers to the generating plants, but also is used by others in carrying on business incidental to the operation of the power system. It was recognized that an investigation of the carrier system should include the characteristics of the extensions because they are a part of the complete communication system. The extensions to the dispatcher's office includes both open wire line and cable. Audio frequency characteristics of one of these extensions are shown in curve *A* of Figure 10. The extension from the dispatcher's office to the general offices is by submarine cable across the San Francisco Bay. The audio frequency characteristics of this cable are shown in curve B of Figure 10. There is in this cable a rapid cutoff of the frequencies above 2,500 cycles. The measured attenuations at 1,000 cycles were approximately two decibels for curve *A* and nine decibels for curve *B*, but the curves have been plotted through zero at 1,000 cycles for a better comparison of their shapes.

Listening tests also were made to compare various types of telephone instruments, and although no quantitative data were obtained, it was apparent that there was a large difference in the performance of different instruments, and that improvements could be made in some cases by providing modern high quality telephone instruments for use with the carrier equipment.

Conclusions

Tests such as those described in this paper provide information to evaluate properly the performance of the various parts of a carrier telephone system, and to make adjustments for optimum performance. Specifically it can be concluded that

1. Care in the adjustment of carrier current communication equipment will provide the maximum in performance, and adjustments cannot be made correctly without the use of proper instruments and a logical procedure.

2. Good audio frequency response is possible by correct adjustment of the equipment, and is particularly important where several sets are to be cascaded by means of audio coupling. Audio frequency response may be limited by characteristics of the line tuning equipment. In practice, a compromise between audio frequency response and signal-to-noise ratio must be accepted.

3. Frequency modulation equipment is capable of maintaining a high signal-to-noise ratio with large variations of carrier signal.

4. The frequency modulation system provides a definite advantage over amplitude modulation equipment insofar as signal-to-noise ratio is concerned, which permits successful carrier current operation on lines with noise levels so high as to preclude the use of conventional amplitude modulated equipment.

5. In frequency modulation, as in any carrier system, it is important to maintain a high level of modulation in order to realize the maximum signal-to-noise ratio.

6. Audio extensions from the carrier terminal equipment enter into the over-all consideration of the system, and can be limiting factors in the performance of a system.

Data such as those collected in the tests described are also invaluable in maintenance work on the system. The actual performance of the equipment can be compared with the test results from time to time, and any marked deviations can serve as guides in finding and correcting troubles.

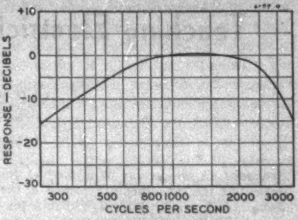

Figure 9. Over-all audio response of three carrier current channels in cascade

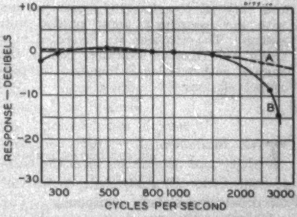

Figure 10. Characteristics of audio frequency extensions to carrier current channels

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Teaching Electricity and Magnetism

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Synopsis: A discussion of a method for presenting the fundamental relations of the science of electricity and magnetism is presented including a brief description of the various absolute unit systems. Most of the fundamental relations of electricity and magnetism are presented in general form. The relations are called general in the sense that they are complete and correct for use with any consistent unrationalized unit system. The relations are presented in a logical and rigorous sequence, that is, the basis of each relation is found in the preceding relations.

THE fundamental relations of electricity and magnetism form the basis of analysis for all analytical problems of electrical circuits and equipment. The primary objective of the formal training in engineering is to develop skill in engineering analysis. Thus it is very important that the electrical engineer have excellent training in his major basis for analysis, electricity and magnetism. Skill in engineering analysis demands an understanding of the field of application and limitations of basic relations which can be gained only through a well-organized study of the science involved.

The purpose of this paper is to present the author's approach to such study and training in electricity and magnetism. The organization presented has evolved through many revisions while presenting the material to sophomore electrical engineering students. It was developed on the basis of maintaining sequential rigor at the sophomore level of comprehension. Electrostatic field theory is developed toward the end of the organization instead of at the beginning because the average student has little background in this field and can gain essential background through the study of circuits and magnetic fields. For example, in accordance with the above principles potential difference is defined as work per unit charge instead of the line integral of field intensity.

Experimental, Defined, and Derived Relations, and Generalizations

The classification of the relations or formulas of electricity and magnetism into experimental, defined, or derived relations and generalizations was suggested to the writer by Professor Edward Bennett and forms an essential part of the organization of the text of which Professor Bennett is the coauthor.[1]

To comprehend fully the significance and limitations of each of the relations, formulas, and rules which make up the body of a science it is necessary to understand clearly the basis upon which each was obtained. All of the relations of electricity and magnetism can be classified under one or another of the four headings. The headings are so descriptive as to hardly need further exposition.

Experimental relations are those which are determined directly by experiment, such as Coulomb's law of force between charges and Ampere's law of force between parallel wires. Relations which are termed experimental were usually so determined historically but this does not imply that all relations which were originally determined experimentally should be so termed. For example, Joule's law concerning power dissipation in a resistor is considered derived in this presentation. It is also important to understand the actual experimental background for experimental relations because they must not be extrapolated too far beyond the experimental evidence on which they are based. Students of electronics are well aware of the limitations of $F = ma$ in cathode ray tube theory. For a presentation of original experimental procedures see Magie[2], and for a discussion of the limitations of experimental relations see Bridgman.[3]

Defined relations include all those which define units, conventions, and physical quantities. It is especially important to understand which quantities or relations are defined because a defined quantity must not be given properties beyond those involved in the definition. Failure to understand that the magnetic field is a defined quantity often leads to fruitless discussions concerning the mechanism of induced voltage. Also it is important that a given quantity be defined in one way only lest inconsistencies appear in succeeding analysis.

Derived relations are determined analytically usually through the combining of experimental and defined relations. Again familiarity with their source is important because all derived relations are circumscribed by the limitations of the experimental and defined relations upon which they are based.

Generalizations or postulates are relations which are assumed to be universally true because no exception to them has ever been observed. The law of conservation of energy is an example. Here again the engineer must understand that the law of conservation of energy based on the Macro measurements of engineering cannot be extrapolated into the realm of atomic physics.

Absolute Measure

A clear understanding of the origin and structure of unit systems is essential in engineering analysis. All electrical unit systems are based on absolute measure, that is, measurement based on the fundamental units of length, mass, time, and a fourth arbitrarily assigned quantity called the permeability of free space. Gauss laid the foundation for absolute measure in the 1830's by measuring the earth's magnetic field in terms of the moment of inertia and period of oscillation of one magnetized needle and of the deflection of a second magnetized needle in the vicinity of the first.[3] The pole strengths of both magnets cancel out of the equations thus giving the strength of the earth's field in terms of length, mass, time, and the permeability of free space tacitly assumed as unity in Gauss' experiments. Thus it was possible to duplicate magnetic measurements without a standard magnetic needle. All electrical units are based on similar measurements. The actual experimental work is very complicated involving difficult adjustments and corrections, but the basic academic definitions of the processes are quite simple as will be shown later.

THE CGS ELECTROMAGNETIC SYSTEM

The cgs electromagnetic system is based on the centimeter, gram, second, and the permeability of free space taken as unity. It is the standard unit system in the field of physics and is the basis of the practical system of electrical units.

THE INTERNATIONAL OR PRACTICAL SYSTEM

The early telegraph engineers did not find the cgs electromagnetic units a con-

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venient size so they developed the practical system whose units differed from the cgs electromagnetic units by various multiples of ten. However, the practical system embraced only circuit phenomena and was not extended to cover mechanics, or magnetic or electrostatic field theory. This was perhaps the greatest source of misunderstanding and confusion in the study of electricity and magnetism since any analysis involving mechanical force or electric or magnetic fields and electric circuits involved borrowing units from other unit systems and resulted in innumerable multiplying factors. Also, no relation could be considered complete without a statement of the particular assortment of units for which it was designed. Giorgi provided the solution in 1903 by the invention of the mks unit system. The system was adopted as the international standard in 1935.

THE MKS SYSTEM

Giorgi's invention of the mks system consisted of extending the practical system into a complete absolute system embracing all of mechanics and electric and magnetic field theory and at the same time retaining all of the circuit units of the practical system. This was done by choosing units of length and mass of such size as to make the unit of mechanical work equal to the unit of work in the practical system (the joule) and by assigning a numerical value to the permeability of free space ($\mu_v = 10^{-7}$) which would result in a unit of current equivalent to the ampere. These ideas will be expanded in the process of setting up the fundamental relations.

GENERAL ELECTRODYNAMIC EQUATIONS

The foregoing discussion indicates the possibility of writing a complete set of fundamental relations in electricity and magnetism with no need for supplementary statements concerning the appropriate units. Units of the mks system are inserted on the right of a relation and the answer is known to be in units of the same system. It may not be quite as evident, however, that these same expressions serve equally well in any other consistent unit system. That is, the identical relations apply equally well in the mks, cgs electromagnetic, and cgs electrostatic systems.

With this approach, beginning students show equal facility in solving problems in any of the three unit systems mentioned. Often one or the other of the cgs systems will handle a given problem with greater convenience than the mks system. Experience has demonstrated that students can use three or more systems interchangeably without confusion. However, it is imperative that only one system be used in a given problem or relation.

The Fundamental Electrical Quantities

ELECTRIC CHARGE

The development of the fundamental relations presented herein is based on the concept of the electron as a charged particle, that is, charge is considered real and field is considered a condition which is produced in the vicinity of charge. The reverse of this may be a more exact picture of nature but the charged particle concept seems to lend itself more readily to the development of the science of electricity and magnetism.

FORCES ON ELECTRIC CHARGE

The science of electricity and magnetism deals with forces between electric charges at rest and in motion. We know from experiment that electric charges exert forces upon each other because of their positions, their velocities, and their accelerations.[4] The first effect is the basis of electrostatics, the second of generated voltages and magnetic forces, and the third of induced or transformer voltages and radio transmission.

ELECTRIC FIELDS

Even though it is possible in many cases to calculate the forces of electric charge in terms of "action at a distance," that is in terms of the force of one charge upon another, it is almost always simpler and more effective to assume that one charge or group of charges produces a field which in turn exerts a force on other charges in the vicinity. Forces between electric charges resulting from their relative positions are calculated in terms of electrostatic fields, and forces between charges in motion are calculated in terms of magnetic fields. The transmission of radio signals is usually calculated in terms of the electromagnetic field, but the calculations also can be made in terms of "vector potentials" which make no use of the magnetic field concept. The defining relations for the various field quantities are presented in the following section.

The Fundamental Electrodynamic Relations and Quantities

The fundamental electrodynamic relations and quantities are herein developed in a rigorous sequence. The source of each relation whether experimental, defined, derived, or generalization is so indicated. Space limitations prevent giving more than an outline of each derivation, or of stating the complete significance of all terms of all of the relations. The prime purpose in presenting the sequence of quantities and relations is to illustrate the logical manner in which the relations of electricity and magnetism may be developed. The reader should refer to standard text books in the field for details of specific relations. Units given are of the mks unrationalized system.

BASIC MECHANICS

1. l—length, meters (defined)
2. m—mass, kilograms (defined)
3. t—time, seconds (defined)
4. v—velocity, meters per second (defined)
5. a—acceleration, meters per second per second (defined)
6. Newton's law of force

$F = ma$ (experimental)

7. F—force, newtons (defined)

A newton is that force which will give a mass of one kilogram an acceleration of one meter per second per second.

8. W—work, joules (newton-meter) (defined)

$W = Fl$

9. P—power, watts, (joules per second) (defined)

$P = W/t$

ELECTRIC CIRCUIT QUANTITIES

10. Ampere's law of force between parallel wires (experimental)

$$F = \frac{K\mu_r I_1 I_2 l}{d}$$

where μ_r is the relative permeability of the medium between the wires, I_1 and I_2 are the currents in the two wires, l is the length of the wires, d the distance between centers, and K is a constant which determines the unit system. To establish an unrationalized system

$K = 2\mu_v$

where μ_v is called the absolute permeability of free space. It is really a number which determines the size of the electrical units.

11. μ_v—absolute permeability of free space, (defined, fundamental mks unit)

$\mu_v = 10^{-7}$

12. $\mu_r\mu_v$—absolute permeability of a medium (defined)

13. I—electric current, ampere (defined)

Unit current is defined in terms of Ampere's law of force between parallel wires

$$F=\frac{2\mu_r\mu_0 I_1 I_2 l}{d}$$

The ampere is that current which will produce 2 × 10^{-7} newtons of force per meter of length between two parallel wires spaced one meter apart in air.

14. *Q*—electric charge (defined)

When two bodies attract each other with a force other than that of gravitation or repel each other with no material connection, they are said to be charged electrically.

15. Relation between electric current and charge (experimental)

$$Q=It$$

Rowland showed in 1876 that electric current was time rate of redistribution of charge.

16. *Q*—charge, coulomb (defined)

One coulomb of charge passes a given point in a circuit each second when the current is one ampere.

17. Potential difference (defined)

Potential difference is the action which tends to set up a current in an electric circuit. The unit is defined in terms of work per unit charge.

18. *E*, *V*—potential difference, volts (defined)

A potential difference of one volt is said to exist between the points *a* and *b* if one joule of energy is required to carry one coulomb of charge from *a* to *b*. The point *b* is at a higher potential than *a* if positive work is done on a positive charge.

$$\begin{aligned}V_{ba}&=\frac{W}{Q}\\&=-\int_a^b\frac{F}{Q}\cos(F,dl)dl\end{aligned}$$

where V_{ba} is the potential of the point *b* with respect to the point *a*, *F* is the electric force on the test charge, and (*F*, *dl*) is the angle between the direction of the electric force and *dl*.

19. Resistance (defined)

Resistance is that property of a circuit which opposes current flow.

20. Ohms Law (experimental)

$$I=E/R$$

21. *R*—resistance, ohms (defined)

A circuit is said to have a resistance of one ohm if a potential difference of one volt sets up a current of one ampere in it.

22. Resistance of a conductor (experimental)

$$R=\frac{\rho l}{A}$$

where ρ is the resistivity of the conductor material, *l* is the length of the conductor, and *A* its cross-sectional area.

23. *G*—conductance, mhos (defined)

Conductance is defined as the reciprocal of resistance.

24. Power in an electric circuit (derived)

$$P=VI$$

This relation is derived from the definition of potential difference, and of current considered as time rate of redistribution of charge.

25. Power in electric circuits (derived)

$$\begin{aligned}P&=I^2R\\&=V^2/R\end{aligned}$$

Derived from (24) and Ohm's law.

26. Energy in electric circuits (derived)

$$\begin{aligned}W&=VIt\\&=I^2Rt\\&=V^2t/R\end{aligned}$$

27. Law of conservation of charge (generalization)

28. Law of conservation of energy (generalization)

29. Kirchhoff's laws (derived)

$$\Sigma E's=0$$
$$\Sigma I's=0$$

These relations are derived on the basis of conservation of energy and conservation of charge.

30. Resistances in series (derived)

$$R_t=R_1+R_2+\ldots\ldots+R_n$$

31. Resistances in parallel (derived)

$$R_t=\frac{1}{\frac{1}{R_1}+\frac{1}{R_2}+\ldots\ldots+\frac{1}{R_n}}$$

MAGNETISM

Considering Ampere's law (13) let us assume that the current I_1 produces a stress in space which exerts a force on the current I_2. This assumption is the basis of the concept of magnetism. Let us say that the stress sets up a quantity called flux Φ (magnetism) usually represented by lines, and that the number of lines per unit area is flux density *B*.

32. Flux density (defined)

$$F=BIl\sin(B,l)$$

Flux density, a vector quantity, is defined by this relation where (*B*, *l*) is the angle between the direction of the flux lines and the direction of the wire.

33. *B*—flux density, webers per square meter (defined)

The flux density of a field is said to be one weber per square meter if a wire carrying a current of one ampere and perpendicular to the field experiences a force of one newton per meter of length.

34. Φ—flux, webers (defined)

$$\Phi=BA$$

or

$$\Phi\int B\cos(B,n)dA$$

The total flux passing through a plane area is equal to the product of flux density *B* and the area *A* if the flux density is uniform and perpendicular to the area. In case of nonuniform fields the summation must be determined by integration. In the integral, (*B*, *n*) is the angle between the direction of flux density and the direction of the normal to the differential area *dA*.

35. Flux density about a long straight wire (derived)

$$B=\frac{2\mu_r\mu_0 I}{x}$$

where *x* is the distance in meters from the center of the wire.

36. Field intensity (defined)

$$B=\mu_r\mu_0 H$$

Field intensity, a vector quantity, is defined by this relation. It is the agency which produces flux density; that is, the flux density at any point in a field is proportional to and directed along the field intensity vector.

37. *H*—field intensity, pragilberts per meter (defined)

A field intensity of one pragilbert per meter sets up a flux density of 10^{-7} webers per square meter in free space.

38. Field intensity about a long straight wire (derived)

$$H=\frac{2I}{x}$$

39. Magnetomotive force (defined)

$$M=\int H\cos(H,dl)dl$$

Magnetomotive force is defined as the line integral of field intensity. It is the total magnetic action along a given path. (*H*, *dl*) is the angle between the direction of *H* and the differential length of path *dl*.

40. *M*—magnetomotive force, pragilberts (defined)

A magnetomotive force of one pragilbert will set up a field intensity of one pragilbert per meter in a path one meter long.

41. Magnetomotive force about a long

straight wire (derived)

$$M=\int_0^{2\pi d}\frac{2I}{x}\cos(H,dl)dl$$
$$=4\pi I$$

It can be shown that the result is independent of the path of integration.

42. Magnetomotive force of a coil (derived)

$$M=4\pi NI$$

N is the number of turns in the coil which is carrying the current I.

43. Field intensity resulting from a current element (generalization)

$$dH=\frac{I\sin(r,dl)dl}{x^2}$$

where dH is the field intensity set up at a distance x from a differential length of conductor dl which is carrying a current I.

This relation must be considered true because no contradictory evidence has been discovered.

44. Field intensity about a wire of finite length (derived)

$$H=\frac{I}{r}(\sin\alpha_1+\sin\alpha_2)$$

Set up (43) for particular case and integrate. x is the perpendicular distance from the point at which H is evaluated to the axis of the wire. α_1 and α_2 are the angles between the perpendicular and the two ends of the wire.

45. Field intensity on the axis of a concentrated circular coil

$$H=\frac{2\pi r^2NI}{(r^2+x^2)^{3/2}}$$

where r is the radius of the coil and x is the distance from the plane of the coil.

Note. (44) and (45) become expressions for flux density by multiplying by $\mu_r\mu_v$.

46. Total flux produced in a toroid of cross-sectional area A and mean length l (derived)

$$\Phi=\frac{4\pi NI\mu_r\mu_vA}{l}\text{ (approximately)}$$

Ratio of length to area must be large.

47. Reluctance (defined)

The reluctance of a magnetic path is defined as the ratio of the magnetomotive force across a path to the flux set up in the path.

48. $\mathcal{R}$—reluctance, pragilberts per weber (defined)

Definition follows from name of unit.

49. Reluctance of a path (derived)

$$\mathcal{R}=\frac{l}{\mu_r\mu_vA}$$

GENERATED AND INDUCED VOLTAGES

50. Electromagnetic induction (experimental)

Faraday discovered that a voltage is generated whenever a conductor is moved across a magnetic field or whenever there is a change in the number of flux lines linking a coil. The first is a case of forces between moving charges and the second of forces between accelerated charges.

51. Voltage generated by a moving conductor (derived)

$$V=Blv\sin(B,v)$$

The derivation is made in terms of an elementary generator and the law of conservation of energy. Wire is considered moving perpendicular to its axis v is the velocity of the conductor and (B,v) is the angle between the direction of the flux and the velocity.

52. Flux linkages, weber-turn (defined)

$$\lambda=N\Phi$$

53. Voltage generated by changing flux linkages (experimental)

$$e=N\frac{d\Phi}{dt}$$

Note. The lower case letter e is used in this and subsequent relations to indicate instantaneous values. Instantaneous current is indicated in a corresponding manner.

54. Inductance (defined)

Inductance is that property of a circuit which tends to prevent a change in current.

55. L—inductance, henries (defined)

$$L=e/(di/dt)$$

A circuit is said to have an inductance of one henry if current changing at the rate of one ampere per second induces an emf of one volt.

56. Inductance in terms of flux linkages per ampere (derived)

$$L=N(d\Phi/di)$$

57. Inductance in terms of circuit dimensions (derived)

$$L=\frac{4\pi\mu_r\mu_vN^2A}{l}\text{ (approximately)}$$

Derived from (56) and (46).

58. Inductance of two long straight wires.

Distance d between centers, wire radius r, and length of circuit l (derived)

$$L=l\left(4\mu_v\log_e\frac{d}{r}+\mu_r\mu_v\right)$$

59. L_m—mutual inductance (defined)

$$L_m=e_1/(di_2/dt)$$

60. k—coefficient of coupling of two coils (defined)

$$k=k_1k_2$$

where $k_1=\Phi_{12}/\Phi_1$ and $k_2=\Phi_{21}/\Phi_2$. Φ_{12} is the part of flux Φ_1 which also links coil 2 and Φ_{21} is the part of flux Φ_2 which also links coil 1.

61. Mutual inductance in terms of self inductance (derived)

$$L_m=k\sqrt{L_1L_2}$$

62. Energy stored in inductive circuits (derived)

$$W=\frac{LI^2}{2}$$
$$=\frac{N\Phi I}{2}$$

63. Energy stored per unit volume in magnetic fields (derived)

$$\frac{W}{v}=\frac{\mu_r\mu_vH^2}{8\pi}$$
$$=\frac{BH}{8\pi}$$
$$=\frac{B^2}{8\pi\mu_r\mu_v}$$

where v is the volume in cubic meters.

64. Pull of an electromagnet (derived)

$$F=\frac{B^2A}{8\pi\mu_v}$$

ELECTROSTATICS

The following relations are concerned with forces experienced by electric charges resulting from their relative positions.

65. Coulomb's law of force between charges (experimental)

$$F=\frac{Q_1Q_2}{\epsilon_r\epsilon_vx^2}$$

where ϵ_r is the relative permittivity of the medium between the charges, also known as the dielectric constant. ϵ_v is the absolute permittivity of free space or more exactly a number the magnitude of which is determined by the unit system. (In an

Table I

Unit System	Permeability of Free Space	Permittivity of Free Space
Mks unrationalized...	10^{-7}	..1.113×10^{-10}
Cgs electromagnetic...	1	..1.113×10^{-21}
Cgs electrostatic......	1.113×10^{-21}..	1
Mks subrationalized..	$4\pi10^{-7}$	..8.854×10^{-12}

electrostatic system it is chosen arbitrarily and determines the size of the electrical units.)

66. ϵ_v—permittivity of free space (experimentally determined constant)

$$\epsilon_v = 1.113 \times 10^{-10}$$

The value may be determined by placing a known charge Q_1 at a given distance from another known charge Q_2 and measuring the force.

67. $\mathcal{E}$—Field intensity (electrostatic) (defined)

$$\mathcal{E} = \frac{F}{Q}$$

Similar to the case of parallel wires, let us assume that one charge produces a condition in space which exerts a force on the other charge. The condition is called field intensity and is defined by the foregoing relation.

68. $\mathcal{E}$—field intensity, newtons per coulomb (defined)

Unit field intensity exerts a force of one newton per coulomb on a point charge.

69. Field intensity in terms of space rate of change of potential, volts per meter (derived)

$$\mathcal{E} \cos (\mathcal{E}, dl) = \frac{-dE}{dl}$$

This relation is derived from the defining relations of field intensity and potential difference.

70. Field intensity about a point charge (derived)

$$\mathcal{E} = \frac{Q}{\epsilon_r \epsilon_v x^2}$$

71. D—electrostatic displacement density or flux density, pracoulombs per square meter (defined)

$$D = \epsilon_r \epsilon_v \mathcal{E}$$

Displacement density is set up by field intensity and is defined by the foregoing relation.

72. ψ—electrostatic displacement or flux, pracoulombs (defined)

$$\psi = DA$$

or

$$\psi = \int D \cos (D, n) dA$$

where n is the direction of the normal to the surface. Also see 34.

73. Displacement from a point charge (derived)

$$\psi = 4\pi Q$$

74. Displacement from any charge (derived)

$$\psi = 4\pi Q$$

from Gauss' theorem.

75. Capacitor (defined)

If a potential difference is applied to two conductors separated by an insulator, charge is observed to flow from one plate to the other. This arrangement is called a capacitor.

76. C—capacitance, farads (defined)

$$C = Q/V$$

A capacitor is said to have a capacitance of one farad if one coulomb is transferred from one plate to the other by a potential difference of one volt.

77. Capacitance of a parallel plate capacitor (derived)

$$C = \frac{\epsilon_r \epsilon_v A}{4\pi x}$$

78. Capacitance of a concentric cylinder capacitor (derived)

$$C = \epsilon_r \epsilon_v l/2 \log_e (r_2/r_1)$$

where l is the length of the cylinder, r_2 the radius of the outer cylinder, and r_1 the radius of the inner cylinder.

79. Energy stored in a capacitor (derived)

$$W = {}^1/_2 CV^2$$

80. Energy stored per unit volume in an electric field (derived)

$$\frac{W}{v} = \frac{\epsilon_r \epsilon_v \mathcal{E}^2}{8\pi}$$

$$= \frac{D^2}{8\pi \epsilon_r \epsilon_v}$$

$$= \frac{\mathcal{E} D}{8\pi}$$

81. Force between parallel plates (derived)

$$F = \frac{2\pi\sigma^2 A}{\epsilon_r \epsilon_v}$$

σ = charge density.

82. Capacitor current (derived)

$$i = c(de/dt)$$

83. Capacitor voltage (derived)

$$e = \frac{1}{C} \int i \, dt$$

Values of Permeability and Permittivity

Table I gives the values of absolute permeability and absolute permittivity in the four common unit systems. In the cgs electrostatic system the permittivity is assigned the value unity by definition and the value for permeability is experimentally determined. The other three, being electromagnetic systems, have the values of permeability assigned arbitrarily and the permittivity values are experimentally determined.

EXAMPLES

The application of various unit systems to the fundamental relations may be illustrated by a few simple examples.

Example 1. Calculate the force per meter of length between two long straight wires spaced 10 centimeters apart and carrying 500 amperes.

Mks (unrationalized)

$$F = \frac{2\mu_r \mu_v I_1 I_2}{r}$$

$$= \frac{2 \times 1 \times 10^{-7} \times 500 \times 500 \times 1}{0.1}$$

$$= 0.5 \text{ newton}$$

Cgs electromagnetic

$$F = \frac{2 \times 1 \times 1 \times 50 \times 50 \times 100}{10}$$

$$= 50{,}000 \text{ dynes}$$

Cgs electrostatic

$$F = \frac{2 \times 1 \times 1.113 \times 10^{-21} \times 500 \times 3 \times 10^9 \times 500 \times 3 \times 10^9}{10}$$

$$= 50{,}000 \text{ dynes}$$

The solutions check since one newton equals 10^5 dynes. The expression does not apply to the rationalized system.

Example 2. Calculate the flux density at 10 centimeters from a long straight wire carrying 50 amperes.

Mks unrationalized

$$F = \frac{2\mu_r \mu_v I}{r}$$

$$= \frac{2 \times 1 \times 10^{-7} \times 50}{0.1}$$

$$= 1 \times 10^{-4} \text{ weber per square meter}$$

Cgs electromagnetic

$$F = \frac{2 \times 1 \times 1 \times 5}{10}$$

$$= 1 \text{ gauss}$$

$$= 1 \text{ maxwell per square centimeter}$$

The calculation can be made similarly for the electrostatic system but there is no

Table II

Quantity	Replace	By
Permeability	μ_v	$\mu_{rv}/4\pi$
Permittivity	ϵ_v	$4\pi\epsilon_{rv}$
Magnetomotive force	M	$4\pi M_r$
Field intensity	H	$4\pi H_r$
Displacement density	D	$4\pi D_r$
Displacement	ψ	$4\pi\psi_r$
Reluctance	R	$4\pi R_r$

accepted name for magnetic flux density in the electrostatic system. The results check again since gausses × 10^{-4} equals webers per square meter.

Example 3. Calculate the capacitance of a parallel plate capacitor consisting of two plates 50 centimeters square and separated by one centimeter. Neglect fringing.

Mks unrationalized

$$C=\frac{\epsilon_r\epsilon_0 A}{4\pi d}$$

$$=\frac{1\times1.113\times10^{-10}\times0.25}{4\pi\times0.01}$$

$$=2.22\times10^{-10}\text{ farads}$$

Cgs electrostatic

$$C=\frac{1\times1\times50^2}{4\pi\times1}$$

$$=198.8\text{ stat-farads}$$

Cgs electromagnetic

$$C=\frac{1\times1.113\times10^{-21}\times50^2}{4\pi\times1}$$

$$=2.22\times10^{-19}\text{ abfarads}$$

See Hudson[5] for a complete list of conversion factors.

Fundamental Rationalized Relations

The fundamental relations can be changed to the rationalized form by making the substitutions indicated in Table II wherever any of the quantities listed appeared in the relations.

The question of the choice between the rationalized and unrationalized systems is not the subject of this paper.

Unit Magnetic Pole

The unit magnetic pole has not been used in connection with the organization because the approach to magnetism through forces on current carrying conductors seems more satisfactory to the writer. Too often the student is introduced to electrical and magnetic quantities through the concept of impossible experiments with unit poles. However, it is granted that the concept may be useful in certain field analyses and there is no objection to defining a unit pole for this purpose just as we have defined magnetic fields for the purpose of simplifying the calculation of forces between electric charges in motion.

Summary

Most of the fundamental relations in electricity and magnetism are presented in general form. The relations are called general in the sense that they are complete and correct for use with any consistent unrationalized unit system. The basis of each relation is clearly indicated and the relations appear in a logical sequence of development. That is, the basis of each succeeding relation is found in the preceeding relations. Such an organization presents the beginning student with the means of acquiring a fundamental understanding of the science of electricity and magnetism, and also forms his basic training in engineering analysis.

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150,000 Horsepower Applied to Aeronautical Research

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ASSOCIATE AIEE

BEHIND THE SCENES of the rapid development of aeronautics in this country during recent years are the aeronautical research laboratories. In addition to their part in the development of aeronautics, these laboratories are of interest because of the variety of electric devices they employ and the surprising magnitude of their electric power loads. One of the largest aeronautical laboratories in terms of electric power is the Ames Aeronautical Laboratory of the National Advisory Committee for Aeronautics, located at Moffett Field, Cal. This laboratory, on which construction was started in 1939, has developed rapidly to the point where it is now the largest single connected load on the Pacific Gas and Electric Company's system. With completion of wind tunnels now under construction this laboratory will have over 150,000 connected horsepower.

The Ames Aeronautical Laboratory is one of three laboratories of the National Advisory Committee for Aeronautics (NACA) the other two laboratories being at Langley Field, Va., and Cleveland, Ohio. The National Advisory Committee for Aeronautics consists of 15 members appointed by the President and serving without pay. This committee serves in an advisory capacity and is charged with the responsibility of studying the fundamental problems of flight with a view to their practical solution. A large supporting staff of full-time employees under the director of aeronautical research operates the facilities just mentioned, and carries out all types of appropriate basic studies of interest to not only the military services and the industry but also to the air lines and private flyers. The Ames laboratory includes several of this country's most up-to-date and highest powered wind tunnels, as well as a highly developed flight research department.

Wind Tunnels

The largest users of power at the Ames laboratory are the wind tunnel fan drives. From the standpoint of control requirements and novel applications of electric machines they are also among the most interesting. A wind tunnel is simply a device for circulating air past an airplane or model of an airplane to simulate the conditions which occur in flight. A schematic sketch of a typical wind tunnel is shown in Figure 1. The propeller, or fan, circulates air around the continuous closed air passage in the direction shown. The air speed is increased to a maximum at the test section, where the model is

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The author acknowledges the assistance of the General Electric Company, the Westinghouse Electric Corporation, and the Allis-Chalmers Manufacturing Company by supplying the detail design of the various drive systems and for their ingenuity and co-operativeness in recommending suitable types of equipment to meet the requirements of the Ames Aeronautical Laboratory.

mounted, by the nozzle-like shape of the air passage at the approach to the test section. Beyond the test section the passage gradually is expanded again to reduce the air velocity and minimize friction losses.

The airplane or model undergoing tests is supported on a balance system so that the forces of the wind on the model may be measured to determine its lift, drag (resistance to forward motion), pitching moments, and so forth. In addition to the over-all forces on the model, forces on individual components may be measured by suitable techniques such as pressure-distribution measurements or by special measuring devices built into the model. The wind tunnel provides a simple device in which controlled test conditions are readily obtainable, and the effects of changes to the model may be studied quickly, accurately, and without the hazards of experimental flight testing. Figure 2 shows a powered model mounted in the 7-by-10-foot wind tunnel.

The power required to operate a wind tunnel ranges from a few horsepower up to many thousands of horsepower, depending upon the size and the air speed produced. The ideal wind tunnel would test full size airplanes at maximum flying speeds. However, such a wind tunnel would require over 400,000 horsepower; consequently, all wind tunnels to date have represented compromises, either as to size or maximum air speed, or both. In the early days of aeronautics much valuable information was obtained from small low speed wind tunnels of a few horsepower. As progress in aeronautics has demanded more and more attention to details the scale-effect errors of such small low speed tunnels no longer can be tolerated. Moreover, as the speed of flight approaches or exceeds the speed of sound (approximately 765 miles per hour at sea level) important phenomena caused by compressibility of the air occur which cannot be simulated in a wind tunnel except by operation at the same high speeds. Thus one finds at the Ames laboratory, for example, the following wind tunnels:

1. The 40-by-80-foot full scale wind tunnel with a test section 80 feet wide by 40 feet high capable of testing actual full size airplanes up to about 70-foot wing span at air speeds up to 250 miles per hour.
2. The 16-foot high speed wind tunnel (16-foot-diameter test section) which tests relatively large models up to about 90 per cent of the speed of sound.
3. The 1-by-3-foot supersonic wind tunnel which tests small models at speeds equivalent to 1,600 miles per hour at sea level.

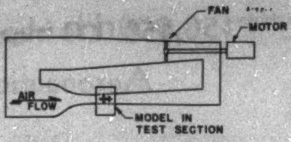


Figure 1. Schematic sketch of the essential features of a wind tunnel

These tunnels require, respectively, 36,000, 27,000, and 10,000 horsepower. Four other wind tunnels of various particular characteristics make a total connected wind-tunnel-fan horsepower of 87,950. An additional 50,000-horsepower supersonic wind tunnel now under construction, plus the miscellaneous laboratory equipment, will bring the total connected horsepower of the Ames laboratory to well over 150,000 horsepower. Figure 3 shows the 27,000-horsepower 16-foot high-speed wind tunnel.

The requirements for the drive system for a wind tunnel are quite exacting in that they usually require adjustable speed over a very wide range and steady speed at any given speed setting, while at the same time often requiring tremendous motor sizes. Operation is intermittent and the equipment must withstand frequent starting, stopping, and speed changing. In large sizes, power factor of the drive system may be important, and starting surges or surges during speed changing may be a critical factor for large wind tunnels.

Of the various types of adjustable speed drives available, the d-c adjustable voltage system (in which the propeller is driven by a d-c motor supplied by a separate generator, and the speed control is obtained by varying the voltage of the generator) has been almost ideal for these applications because of its simplicity of control, wide speed range without discontinuities, and adaptability to automatic control. This system has been used for most small and medium sized wind tunnels. Various other types of systems have been used for larger tunnels, either in an attempt to decrease the cost over that of the d-c adjustable-voltage system or to avoid problems which arise in applying the d-c system in very large sizes. In recent years what is sometimes called the modified Kraemer system has been used very successfully on several large wind tunnel drives, including the Ames 36,000- and 27,000-horsepower tunnels. In this system the main propeller motor is a wound-rotor induction motor, and speed control is obtained by impressing a voltage of controllable frequency on the rotor of the induction motor.

Modified Kraemer Adjustable Speed Drive

To understand the operation of the modified Kraemer system, it might be well to consider first the action of the wound-rotor induction motor with conventional speed control by means of secondary resistance. The voltages and power in the rotor circuit are indicated in Figure 4a. For ease in understanding, the rotor circuit is shown in Figure 4 as though single phase, and, in determining components of power, internal losses are neglected. As the resistance connected across the rotor is increased, the rotor must slip a greater amount in respect to the rotating stator flux in order to generate sufficient voltage in the rotor circuit. This circulates sufficient current in the rotor to develop the necessary torque, resulting in a decrease in speed. By varying the rotor resistance a range of speeds may be obtained. The current and voltage in the rotor circuit will be alternating at a frequency corresponding to the slip between the rotor and the stator flux. As far as the rotor circuit is concerned, the motor acts like a generator and, neglecting motor internal losses, induction motor theory shows that

Figure 2. A powered model mounted in the 7-by-10-foot wind tunnel at the NACA Ames Aeronautical Laboratory, Moffett Field, Calif.

Figure 3. The 27,000-horsepower 16-foot high-speed wind tunnel at the NACA Ames Aeronautical Laboratory

an amount of power is dissipated in the resistor equal to Ps, where P is the power input to the motor and s is the slip in fraction of synchronous speed. Thus power is wasted, resulting in a low efficiency which cannot exceed $(P - Ps)/P = 1 - s =$ speed in fraction of synchronous speed.

If it were possible to generate a voltage whose frequency would at all times be equal to the slip frequency of the motor and in the proper phase relationship, it would be possible to substitute this generated voltage in place of the voltage drop across the resistor. An alternator may be considered as a source for such a voltage. At first glance it would seem impracticable, in replacing the resistance by an alternator, to keep the generated frequency the same as the rotor frequency so that the generated voltage at all times would be exactly opposite in phase to the voltage induced in the rotor circuit. However, upon studying the arrangement further, it will be found that any tendency for the alternator voltage to change in frequency or phase in respect to the rotor voltage will result in a circulating current which will produce a synchronizing torque tending to prevent the two from getting out of phase. This action is very similar to that which occurs in the case of a synchronous motor which, within the limits of its pull-out torque, always will remain with its internal generated voltage in phase opposition to the line voltage. If the frequency of the alternator is decreased by decreasing the alternator speed, the synchronizing current induced between the alternator and the induction motor rotor will increase the motor torque causing the rotor to speed up, thus decreasing the slip frequency and permitting the two to remain in step. Conversely, if the alternator frequency is increased, the resulting circulating current will cause the induction motor to decrease in speed so that the increased slip frequency still matches the alternator frequency. Thus, the speed of the induction motor may be controlled over the operating range by varying the speed of the alternator connected to the rotor. Similarly, as the main motor is loaded and tends to slow down, the corresponding phase shift between the rotor voltage and the alternator voltage causes a circulating current to flow which produces the necessary torque to support the load. The same circulating current flowing through the alternator tends to speed up the alternator, causing it to act as a motor. The flow of power is shown in Figure 4b.

The speed of the alternator can be adjusted most readily by driving it by a d-c motor supplied from a d-c generator with adjustable voltage speed control. The complete system is outlined in Figure 5 for the case of a single fan motor.

The alternator must be of sufficient size to handle the full load rotor current of the induction motor, and must be able to produce, at the primary frequency, a voltage equal to the rotor voltage of the induction motor at standstill; although, with a fan load, the conditions of maximum current and maximum voltage never occur simultaneously. The power which must be handled by the motor-generator sets under any particular condition of load and speed must be equal to the power in the induction motor rotor circuit which otherwise would have been dissipated in a secondary resistance. This power, as previously stated, is equal to Ps, or, in terms of the motor output, the rotor power is $P_o[s/(1\text{-}s)]$, where P_o is the power output. For a fan load varying as the cube of the speed, such as is encountered in the usual wind tunnel drive, the rotor power reaches a maximum value at two-thirds synchronous speed equal to approximately one-seventh of the motor horsepower rating. Thus, it may be seen that most of the auxiliary machines are much smaller in size than would be the motor-generator set for a straight d-c adjustable-voltage system.

This modified Kraemer system is considerably more complicated in the number of machines involved and the complexity of the control than a straight d-c adjustable-voltage system. In fact, it involves, in addition to the wound-rotor motor, a complete adjustable-voltage system for secondary frequency control. The advantage of the modified Kraemer system on extra large drives lies in the reduced size of the d-c machines, and the fact that the equipment may be started with less disturbance to the power system than in the case of starting the adjustable-voltage d-c system. This system has essentially all the desirable control characteristics of the d-c adjustable-voltage system, including ready control of power factor and good efficiency.

In the usual starting sequence the constant speed motor–generator set is first started. This set is quite small compared with the rating of the main motor and offers no difficulties from the standpoint of starting surge. The variable speed set then is brought up to speed with the alternator excited until the alternator frequency is 60 cycles. Since

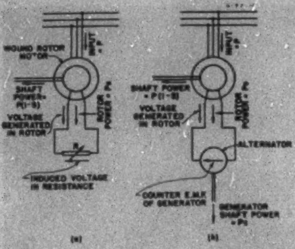

WOUND ROTOR MOTOR
INPUT = P
SHAFT POWER = P(1-S)
ROTOR POWER = Ps
VOLTAGE GENERATED IN ROTOR
R
INDUCED VOLTAGE IN RESISTANCE
(a)
INPUT = P
SHAFT POWER = P(1-S)
ROTOR POWER = Ps
VOLTAGE GENERATED IN ROTOR
ALTERNATOR
COUNTER E.M.F. OF GENERATOR
GENERATOR SHAFT POWER = Ps
(b)

Figure 4. Voltage and power relations in the circuit of a wound-rotor induction motor

(a.) With resistance in secondary circuit
(b.) With counter electromotive force of an alternator in secondary circuit

the alternator is connected to the main motor rotor at all times, a voltage appears on the primary of the motor. By adjusting the alternator speed the motor stator voltage may be synchronized with the line voltage and the motor primary breaker closed with no attendant power disturbance. The motor then may be caused to rotate at any desired speed by reducing the speed of the alternator. With this system a 27,000-horsepower motor can be brought onto the line with no more disturbance than that incurred in starting a 5,000-horsepower synchronous motor.

While exciting the main motor from the secondary side prior to synchronizing, the d-c machine of the constant speed set is generating, driving the variable speed set. After synchronizing, when the alternator frequency is reduced and the main motor starts rotating and picking up load, the flow of power is reversed. Thus the various machines operate sometimes as generators and at other times as motors, and the terms generator or motor are meaningless.

Figure 5 shows a single fan motor. The modified Kraemer system functions equally well if the load is divided among two or more motors with their rotors connected to a single alternator. In fact the 36,000-horsepower wind tunnel, because of its tremendous size, actually uses six 6,000-horsepower motors and fans arranged aerodynamically in parallel in a double row of three above three (Figure 6). The motors are mounted in the air stream, housed in the faired enclosures immediately in front of the propellers. Since all motors are connected to a common frequency on the stator side and to another common frequency on the rotor side, they operate in exact synchronism with each other. In the 16-foot wind tunnel the 27,000 horsepower is divided between two 13,500-horsepower motors driving separate counterrotating fans in tandem (Figure 7). The two motors are mounted back to back in the steel enclosing drum between the two fans.

A doubly fed induction motor is similar to a synchronous motor in that it has a revolving stator field and a separately excited rotor field, the two fields being locked into step by the synchronizing torque. As a synchronous motor tends to oscillate, or hunt, so also does the doubly fed induction motor. A damper winding cannot be used on the induction motor, as on the synchronous motor, because the magnetic flux normally rotates in respect to the rotor at slip frequency and would generate a high voltage in the damper winding even though there were no hunting. Thus the hunting problem demands careful attention on the part of the designer. Where there is a single motor, damping can be incorporated in a damper winding in the variable-frequency alternator. However, if two or more induction motors are connected in parallel to a single alternator, it is possible for the motors to oscillate in respect to each other without affecting the alternator, thus obtaining no help from the damper winding. In the 16-foot wind tunnel with its two motors, the manufacturer avoided independent oscillations between the motors by connecting the variable frequency alternator in series in the rotor circuit between the two motors so that circulating current between the two motors also must pass through the alternator. In the 40-by-80-foot wind tunnel with its six motors there was no alternative to the parallel connection, and hunting was prevented only by careful attention to damping factors in the design.

Figure 8 shows the control motor–generator sets at the 16-foot wind tunnel.

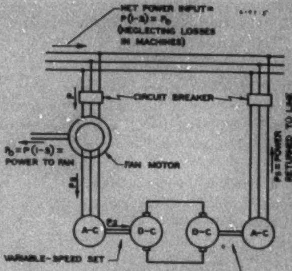

NET POWER INPUT = P(1-s) = P_b (NEGLECTING LOSSES IN MACHINES)
P
CIRCUIT BREAKER
P_b = P(1-s) = POWER TO FAN
FAN MOTOR
Ps
Ps = POWER RETURNED TO LINE
A-C
Ps
D-C
D-C
A-C
VARIABLE-SPEED SET
CONSTANT-SPEED SET

Figure 5. Connection diagram for a complete modified Kraemer system

Two d-c machines are seen on each set because it was impracticable to handle the full power in a single d-c machine. Figure 9 shows a view of the associated switchgear. Modern metal clad switchgear is used throughout, using oil circuit breakers of 500,000 kva interrupting capacity. Controls are semiautomatic so that the wind tunnel operator can devote his full attention to the test under progress. Only such simple controls as a "start-stop" switch and a speed control switch are required to give the operator complete control of the massive wind tunnel drive equipment. When starting a run the operator simply turns the start-stop switch to "start." Without further attention on his part the equipment automatically goes through the sequence of starting the various auxiliary machines and motor–generator sets and automatically synchronizing them with the line. When the main motors are on the line a red light is illuminated on the control panel signalling to the operator that the equipment is on the line ready for him to adjust the speed as needed. The complete start consumes approximately four minutes.

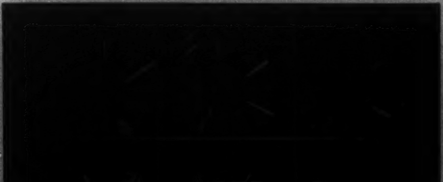


Figure 6. View looking upstream at fans of the NACA 40-by-80-foot wind tunnel—six fans, each driven by a direct - connected 6,000 - horsepower motor

COMBINATION SLIP-REGULATOR AND D-C DRIVE

The 12-foot pressure wind tunnel, with 10,250 fan horsepower, uses to advantage a combination of d-c adjustable-voltage control and wound-rotor induction motor with secondary resistance control. The fan drive unit consists of a 1,250-horsepower d-c motor and a 9,000-horsepower wound-rotor induction motor in tandem. A motor–generator set supplies adjustable voltage direct current for the d-c motor, while a liquid rheostat provides an adjustable resistance for the induction motor rotor circuit. Since the fan load of a wind tunnel varies approximately as the cube of the speed, a tunnel requiring 10,250 horsepower at full speed would require only 1,280 horsepower at one-half speed. Thus the 1,250-horsepower d-c motor provides ample power up to almost half speed. Above half speed the induction motor is energized and its secondary resistance adjusted so that it carries the excess load. At full speed both motors are carrying full load. Sensitive automatic speed regulation is maintained at all speeds through the d-c motor to prevent fluctuations in speed.

SUPERSONIC TUNNEL DRIVES

The wind tunnels discussed thus far all operate at test air speeds below the speed of sound. Wind tunnels designed for operation at test section air speeds in excess of the speed of sound, that is, supersonic wind tunnels, exhibit some quite different characteristics.

1. In the supersonic tunnel the volume of air handled is relatively small and pressure differences relatively high; consequently, the "fan" of a supersonic tunnel has less of the characteristics of an ordinary fan or airplane propeller and is more like a turbine or air compressor. In fact, the 1-by-3-foot supersonic wind tunnel uses commercial centrifugal type air compressors to move the air.
2. At supersonic speeds, an attempt to increase the air speed by speeding up the blower results primarily in an increase in air density at the test section rather than an increase in air speed. Hence air speed control in this class of wind tunnel must be obtained by changing the nozzle contours and the cross sectional area of the test section, and a constant speed motor may be used to drive the blower.

The 10,000 horsepower 1-by-3-foot supersonic wind tunnel utilizes four 2,500-horsepower synchronous motors driving individual compressors. The 50,000-horsepower supersonic wind tunnel now under construction will utilize two 25,000-horsepower wound-rotor induction motors driving an axial-flow compressor. However, wound-rotor motors were selected instead of synchronous motors, not for speed control, but rather from considerations of starting of large machines with high WR^2 load.

Variable Frequency Systems

In conjunction with the wind tunnels another major application of electric power is in the operation of model propellers on models undergoing tests. The rapid increase in the engine power in airplanes in recent years has made the effects of the propeller slipstream and propeller thrust so important an item in the characteristics and control of the airplane that it is no longer possible to obtain sufficiently useful test information on the model without duplicating the action of the propellers. In a full-scale wind tunnel it is usually convenient to use the actual service engine and propeller when testing airplanes. In models, however, it is necessary to provide some other source of propeller power. When a small-scale model is scaled down from the actual airplane, it generally is found that the space available for a motor to provide the motive power for the propeller on the model is much too small for any standard motor. Also, the rotational speed required for the propeller is very high. In scaling down the propeller operation from a full-scale airplane to a small model, it is necessary that the tip speed of the propeller on the model be the same as the tip speed on the full-scale propeller. This necessitates that the propeller speed be increased in proportion to the decrease in its diameter. Thus, if a full-scale propeller operates at 1,500 rpm, the propeller on a 1/10-scale model would have to turn at 15,000 rpm.

Figure 7. View looking upstream at fans of NACA 16-foot high speed wind tunnel at the Ames Aeronautical Laboratory—two fans, each direct connected to a 13,500-horsepower motor

To meet these difficult conditions, the most practical solution has been the use of 3-phase squirrel cage induction motors, since this type of motor is simplest in construction and smallest in size for its horsepower rating. Even then, in order to get the power within the available space, all possible means must be taken to keep size at a minimum. Consequently, the motors are water cooled and are operated at temperatures which would not be tolerated in ordinary motors, but which give a reasonable length of life for such test purposes. A few small motors have even used carbon dioxide from high pressure cylinders as a refrigerant for cooling. As examples of what can be obtained in motors of this type, motors rated 22 horsepower at 12,000 rpm and measuring only $4^1/_2$ inches in diameter by 12 inches long and motors rated 100 horsepower at 10,000 rpm measuring 8 inches in diameter by $16^1/_8$ inches in length are typical of water cooled models. A motor 2 inches in diameter by $7^5/_8$ inches long is available rated 4 horsepower at 18,000 rpm with combined water and carbon dioxide cooling. Figure

Figure 8. Control motor – generator sets for NACA 16-foot wind tunnel drive system at the Ames Aeronautical Laboratory

Figure 9. Switchgear for control of the NACA 16-foot wind-tunnel drive at the Ames Aeronautical Laboratory

10 shows the 4-horsepower and 100-horsepower motors with a conventional 1-horsepower general purpose motor for comparison. These motors for models usually are designed to carry their rated load for approximately 20 minutes, which allows sufficient time for the average test. They have been built in ratings from 2 horsepower at 18,000 rpm to 1,000 horsepower at 2,400 rpm. The aeronautical laboratories continually are attempting to produce smaller and more powerful motors.

On these high-power high-speed motors, bearings are a major problem. Thus far, antifriction bearings have been used exclusively. A major addition to bearing troubles arises from the dissipation of the heat from the rotor of the motor. The stator is readily water-cooled, but there is no way of applying water to the rotor. Consequently, any heat generated in the rotor tends to travel along the shaft and heat the bearings. The point has not been reached where the bearings can be depended upon completely. However, it has been found necessary to provide the bearings with considerable internal looseness of the balls in the races in order to allow the inner race to expand without causing the bearing to become tight.

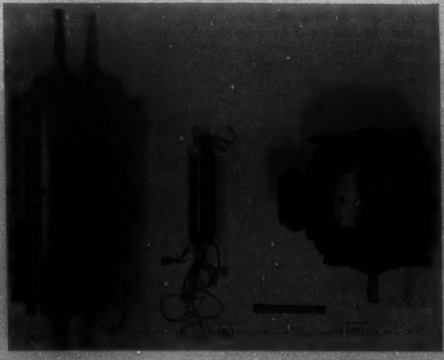
In nearly all tests using model propellers, it is essential to be able to vary the propeller speed over a wide range. Since the motors are of the squirrel cage induction type, the only means available for varying their speed is to change the frequency of the power supply from which they are operated. Thus, a variable frequency power supply is a necessary accessory for power-on tests. These variable frequency power supplies consist of either an alternator or a frequency changer driven at any speed over a wide speed range by a d-c motor supplied from a separate d-c generator with adjustable voltage control. In order to obtain the necessary high speeds with squirrel cage induction motors, it is necessary to operate them on high frequencies. Thus approximately 300 cycles per second are required for 18,000 rpm. There are seven separate variable frequency systems at the Ames laboratory of ratings from 100 kw at 400 cycles to 1,800 kw at 150 cycles. The motors usually are operated at a constant ratio of voltage to frequency because this results in a constant magnetic flux in the motor and gives approximately a constant value of current for a given torque throughout the speed range. However, different values of volts per cycle are required for different motors.

The distribution of the output of each variable frequency set to several points of utilization offers interesting problems. The ordinary a-c power distribution system is a simple parallel system—the voltage and frequency are constant, the circuit is energized at all times, and motors and equipment necessarily are provided with any features required for starting on a full voltage system. Any number of loads can be manipulated independently without appreciably affecting each other. On variable frequency systems, however, only a single load can be served at a given time from one variable frequency set since each load requires individual control of voltage and frequency. Controls must be provided at each point of utilization for varying the voltage and frequency, and interlocking must be provided to insure that only one station has control at any one time, all other loads being disconnected. Variable frequency motors normally are started on reduced frequency and voltage without auxiliary starting equipment. This necessitates interlocking equipment to prevent application of the power unless the voltage and frequency are at low values. These requirements result in rather involved systems when one variable frequency set must serve several different load stations, or when provisions are required for paralleling sets for added capacity. Distribution problems are complicated further by high voltage drop in cables because of high reactance at the high frequencies involved.

The same line of high speed motors and variable frequency supply systems is used to drive blowers, instead of propellers, to simulate the jet from a jet engine. The testing of jet propulsion combinations has introduced a new need for extremely high speed motors in high horsepower ratings.

Power Supply and Distribution System

Electric power is delivered to the laboratory at 110,000 volts from the Pacific Gas and Electric Company's large sub-



A. A 100-horsepower 10,000-rpm water cooled motor

B. A 4-horsepower 18,000-rpm water- and carbon-dioxide-cooled motor

C. A 1-horsepower 1,725-rpm conventional general purpose motor (shown for comparison)

Figure 10. Two typical motors for powered aircraft models shown in comparison with a conventional general purpose motor

station at Newark, Cal. A 50,000-kva and a 5,000-kva transformer bank reduce the voltage to 6,600 volts for distribution to the various loads which all are located relatively close to the substation. For the 50,000-horsepower supersonic tunnel under construction, an additional 62,500-kva transformer bank is being installed.

Owing to the large concentration of power, careful consideration has been required to keep the short-circuit capacity on the 6,600-volt feeders under 500,000 kva. To exceed the 500,000-kva limit would increase the cost of motor-control switchgear greatly as well as increase the probable extent of the damage in case of a fault. To keep within the 500,000-kva limit it was found necessary to avoid interconnecting the 6,600-volt busses of the proposed 62,500-kva bank and the existing 50,000-kva bank and install a small amount of reactance in the feeders to the 50,000-horsepower supersonic tunnel. The machines of the large wind tunnel drive systems make appreciable contributions to the short-circuit capacities on the 6,600-volt busses.

Power is distrubuted at 6,600 volts to the various buildings for small power and lighting. Y-connected 120/208-volt systems, on which 120-volt single phase loads may be distributed on all three phases, are used to advantage in the buildings, where the loads are a combination of lights and 3-phase motors. In a new hangar and shop building of much larger size than previous buildings, a 440-volt system was installed for building distribution and for motor loads with 440- to 120/208-volt dry-type transformers at load centers for lighting and convenience outlets.

Wind tunnels always operate intermittently, and much of the time at less than maximum rated power. Consequently, by staggering operation of several wind tunnels, it is possible to operate under a maximum kilowatt demand limit much less than the total kilowatt rating of the apparatus. In order to make most efficient use of power within the established maximum demand limit without unnecessarily delaying any wind tunnel, a totalizing and telemetering system is being installed which will indicate at all times at each wind tunnel control desk and at a central dispatcher's office the total kilowatt demand for the entire laboratory.

Fault-Current Measuring Device

MARTIN J. LANTZ
ASSOCIATE AIEE

IN electric power system operation numerous problems are encountered in analyzing the results of short circuits and in the physical location of the trouble. Some devices now available for analyzing short circuits in a-c circuits are the oscillograph, the annunciator ammeter, and a magnetic link device requiring rectifiers. The oscillograph is the most accurate device but is expensive. The annunciator ammeter provides a limited amount of data; however, it offers a high electric burden to the circuit and does not have closely graduated steps of reading.

When a short circuit occurs on a power system, the current normally increases immediately in the faulted area. The device that records the phenomena must be very rapid and positive in action. It must not interfere with the electrical characteristics of protective apparatus in the same circuit, and it must be prepared to operate automatically, recording values that may be interpreted later. To be extensively applied, the device must be inexpensive, reasonably accurate, and reliable.

The fault-current measuring device described in this paper meets these requirements.

Description of Device

The fault-current measuring device for use in a power circuit consists of one or more coils of insulated copper wire with magnetic links snugly fitted within them, and a standard high-speed voltage relay with low resistance and normally closed contacts shunting the coils (Figures 1–3). The use of a surge-crest ammeter and a demagnetizing coil is required to read the magnetization of the links and to demagnetize them after energization. The surge-crest ammeter is a special device used principally in lightning investigation for measuring the degree of magnetization of magnetic links. The magnetic link consists of three or four small magnetic laminations enclosed in a small tube. The device may be built with one or more coils to cover the desired current range. It is designed to operate in the secondary circuit of current transformers and is placed in series with the regular current-operated system protective relays or in the metering circuits. The burden of the device is low.

The high-speed by-pass relay is used with the device to minimize the effect of the initial d-c offset that may occur when the magnitude of an inductive alternating current is suddenly increased. In the field test application the by-pass relay was energized from the same direct voltage circuit that energized the tripping circuit of the oil circuit breaker. The by-pass relay which operates as a result of action by the system protective relays then opens its normally closed contacts across the coils one cycle or more prior to the zero point of the interrupted wave, allowing the current to be measured to flow through the coils and magnetize the links. The magnetic links are in the electric circuit when used with the by-pass relay only after the relay is energized and the by-pass contacts are open. Figure 4 is a typical schematic installation of the device. After operation of the device a target shows, the coils are again by-passed, and the by-pass relay locks out to prevent a second operation. A second device may be placed in service automatically if desired.

One device designed and tested covers a current range of approximately 4 to 60 amperes rms alternating current and consists of a 12- and a 56-turn coil in series. Assuming a 600/5 ratio current transformer, a primary current range of approximately 500 to 7,200 amperes is covered.

Theory

The magnetic link is placed in the coil and is magnetized by an alternating current flowing through the coil. This alternately magnetizes the link in one direction and then the other, the magnetism following the link's hysteresis loop. Each half cycle of symmetrical current demagnetizes the magnetism resulting from the previous half cycle and remagnetizes the link in the opposite direction. The electrical principle of the device is based upon the hysteresis loop of the small magnetic link and the interruption of the current at the zero point

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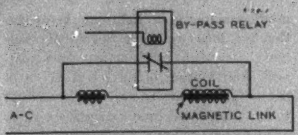


Figure 1. Schematic diagram of basic device for fault-current measurement

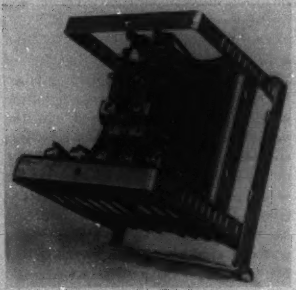


Figure 2. Rear view of preliminary sample of fault device showing relays

Device is mounted in cradle of standard single element protective relay

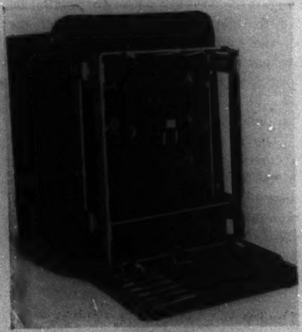


Figure 3. Front view of preliminary sample of fault device showing coils and links

(Figure 5). The residual magnetism retained by the magnetic link is proportional to the peak or crest magnitude of the last half cycle of the current, providing the last two half cycles are approximately equal in magnitude. This magnetism is then measured with a surge-crest ammeter. A calibration curve is obtained by plotting surge-crest ammeter readings against alternating current values. This provides a direct means for determining the last half cycle of alternating current through the coil by interrupting the current at zero and reading the degree of link magnetization by the surge-crest ammeter.

If the current is asymmetrical containing a d-c component, the magnetism of the link is affected and records a value which will not compare with a calculated or measured symmetrical value. The initial surge of direct current may place an amount of magnetism in the link that cannot be wiped out before the circuit is interrupted by the successive half cycles of opposite direction. The by-pass relay effectively eliminates the d-c component from the link as long as it is by-passed and then applies a reduced d-c component to the link, resulting in a more accurate determination of the magnitude of the last half cycle. By placing in series with the coils a 0.2- to 0.3-ohm resistance, the current is diverted to the by-pass relay contacts. This is a precaution against contact resistance and adds only a small burden to the circuit during the time the by-pass contacts are open. The illustrated replica of a test oscillogram demonstrates the action of this by-pass relay (Figure 6).

Use and Advantages

The basic device, which is considered to be the coil and link, will record automatically the magnitude of the last half cycle of an alternating current wave preceding its interruption at current zero by an oil circuit breaker. To obtain data of practical value on power systems with high-speed oil circuit breakers which may not allow time before opening for the d-c component effect to neutralize itself in the link, the by-pass relay is a necessary adjunct to the device. The data obtained by the device will aid greatly in determining the actual current values from short circuits on power systems.

Table I. Summary of Laboratory Tests on Fault-Current Measuring Device

Number of Turns in Coil	Number of Readings	Max Deviation, Per Cent	Average Deviation, Per Cent	Current Range, Amperes
45	6	−10.7	+2.1, −6.3	5.3–17
20	4	+ 8.7	+4.5, −4.2	9.4–38
12	12	+ 9.8	+3.3, −4.8	15.6–61
10	7	−10.6	+5.3, −7.8	18.8–77
5	6	+10.5	+8.7, −5.3	38.5–84

The following is a list of some of the advantages over devices which accomplish similar purposes, such as the automatic oscillograph and the annunciator ammeter:

1. Simplicity. There are no moving parts in the recording portion of the device, therefore no mechanical failures resulting from restrained movement can occur. The by-pass relay is considered as auxiliary to the basic device.

2. Instantaneous operation. There is no mechanical inertia in the recording portion of the device. It requires no mechanical action and is ready at all times for instant record.

3. Durable record. It is desirable to read the device soon after its operation but the reading may be deferred a number of hours.

4. It records a value proportional to the magnitude of alternating current without resorting to initial rectification.

5. Used in conjunction with a by-pass relay to record the last half cycle of a transient asymmetrical wave, the effect of the possible high initial d-c offset resulting from short circuits is materially reduced.

6. The device has a low electric burden and does not jeopardize the operation of protective devices.

7. The device may be automatically locked out of the circuit after operating.

8. The readings obtainable from the device are a continuous graduation of values obtained from a calibration curve allowing more accurate analysis to be made than is possible from a step-reading device such as the annunciator ammeter.

9. The device is inexpensive, making it

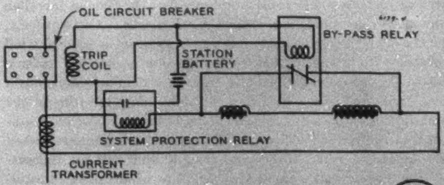


Figure 4. Schematic diagram of typical installation of fault-current device

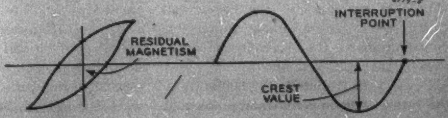


Figure 5. Hysteresis loop and sine wave

possible to distribute them extensively about the system to obtain operational records. As a fault-locating device and indicator of the type of fault, it can be applied to many lines where a more expensive device is not feasible.

Results of Tests

A series of tests using the magnetic link was made by the Bonneville Power Administration's electrical laboratory. These constituted both laboratory and field tests.

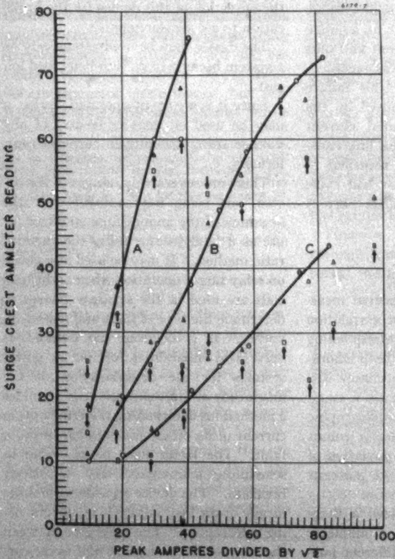
Laboratory Tests

The laboratory tests provided a-c and d-c magnetic link calibrations for various turn coils and tests to determine the effectiveness of the by-pass relay on asymmetrical waves. Data were secured for 5-, 10-, 12-, 20-, and 45-turn coils.

Figure 7. Fault device laboratory test on 5-, 10-, and 20-turn coils

⊙—A-c calibration
△—Link reading short-circuited by relay
⊡—Link reading not short-circuited by relay

Curve	Link	Turns of Coil	Max Deviation, Per Cent	Avg Deviation, Per Cent	Current Range, Amp
A	8	20	8.7	4.5, −4.2	9.4 to 38.5
B	6	10	10.6	5.3, −7.8	18.8 to 77
C	4	5	10.5	8.7, −5.3	38.5 to 84.8



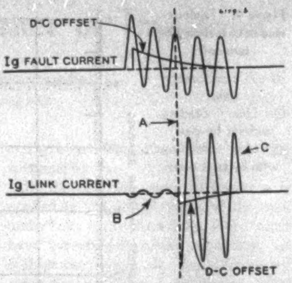


Figure 6. Replica of oscillogram showing results of use of by-pass relay across coil during system test

A—Current through coil when by-pass relay contacts are opened
B—Current through coil when by-pass relay contacts are closed
C—Last one-half cycle current measured by link

The a-c calibration data were obtained by adjusting the current with a variable resistive load. The energized circuit was closed through a circuit breaker into a resistance load, the circuit breaker opening at the zero point of current. Oscillographic records were obtained for all tests so that the results could be correlated. The a-c calibration points are indicated in Figure 5 by the circles. Peak amperes divided by the square root of two are plotted against surge-crest ammeter readings.

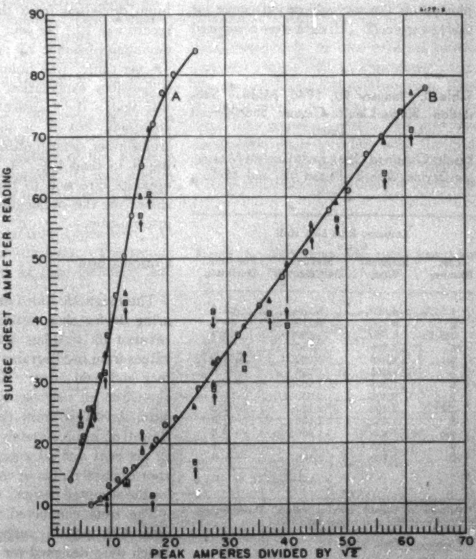
To determine the effect of the by-pass relay on offset currents, the resistance load was replaced with an inductive load. The current was passed through two coils in series, one of which was by-passed for a few cycles. The points in the squares, Figure 7, indicate the link readings containing the effect of d-c offset. The points in the triangles indicate the link readings in the by-passed coil. Shunting most of the d-c offset current around the link causes the link reading to move closer to the calibration curve.

Similar data are shown in Figure 8 for the 12- and 45-turn coils. Since the surge-crest ammeter readings are more accurate above 18 and below 70, these values were used as the limits of reliability.

Figure 8. Fault device laboratory test on 12- and 45-turn coils

⊙—D-c calibration
△—Link reading short-circuited by relay
⊡—Link reading not short-circuited by relay

Curve	Link	Turns of Coil	Max Deviation, Per Cent	Avg Deviation, Per Cent	Current Range, Amp
A	8	45	10	2.1, −6.3	5.3 to 17.2
B	6	12	9.8	3.3, −4.8	5.6 to 61.1



The relationship between actual oscillographic measured current values and current values measured by means of the magnetic link are expressed in Table I as percentages of the link reading.

Theoretically the magnetic link should be magnetized in proportion to the peak value of current for a sinusoidal wave. Tests show the magnetic link d-c calibration curve to lie slightly higher than the corresponding a-c calibration curve which is to be expected after the first half cycle of magnetization.

All tests conducted should give link readings corresponding to the a-c calibration curve. However, the tests conducted in the laboratory corresponded more closely to the d-c calibration curves. This was probably the result of a higher percentage of harmonics introduced by the highly reactive test load used.

Tests in the field corresponded more closely to the a-c calibration curve as expected. The a-c calibration curve was used as the basis for determining the field test currents.

FIELD TESTS

A series of staged system fault tests was made in May 1945, including single-phase-to-ground, two-phase-to-ground, and phase-to-phase faults. A fault-current measuring device with a by-pass relay was installed during the tests. The data resulting from these tests are shown in Figure 9. The test data were plotted on the a-c calibration curve for the 12-turn coil. All data were compared with oscillographic records taken during the test.

Table II. January 20, 1946, Midway Substation Single-Line-to-Ground Short-Circuit Tests

Results Obtained Using Fault-Current Measuring Device With By-Passed 12- and 45-Turn Coils

Test Number	Amperes Rms Last Half Cycle: Magnetic Link	Amperes Rms Last Half Cycle: Magnetic Oscillograph	Per Cent† Deviation
1	4,960	5,250	−5.9
2	4,770	4,770	0
2R	4,860	4,680	+3.7
*3	3,700	5,840	
4R	5,380	5,220	+3.0
5	1,295	1,290	+0.3
6	1,295	1,305	−0.8
7	2,305	2,030	+11.8
7R	2,400	2,170	+9.5
8	1,655	1,660	−0.3
8R	1,655	1,660	−0.3
9	1,645	1,845	−12.1
10	1,260	1,190	+5.6
		Average	−3.9, +4.8

* Failure to record correct value. Possible explanation in the text under field tests.

† $\frac{I_{\text{oscillograph}} - I_{\text{link}}}{I_{\text{link}}}$

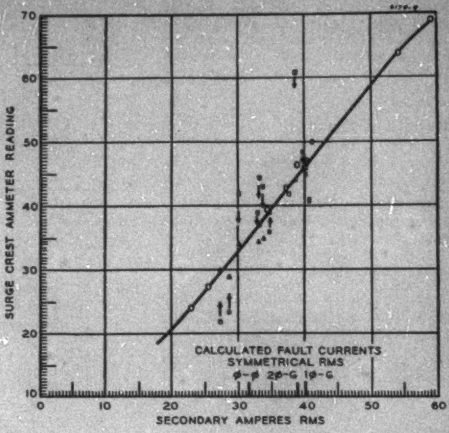


Figure 9. Fault device field test on 12-turn coil

⊙—A-c calibration data

⊡—Link reading without by-pass

△—Link reading with by-pass relay

The oscillograms of the tests show a higher degree of d-c offset present than normally would be expected for system faults. Several tests resulted in a rather large amount of d-c offset still present at the time of interruption.

A second field test was conducted in January 1946, during which thirteen single-phase-to-ground faults were made on the system. The results of this test are summarized in Table II. A maximum deviation from the oscillographic record was 12.1 per cent with an average deviation of minus 3.9 per cent and plus 4.8 per cent. One failure was recorded. A possible explanation for this failure was that a d-c transient caused by the collapsing flux of a saturated current transformer demagnetized the link causing it to read low. A d-c transient of opposite direction to the last half cycle shows on the oscillogram after current interruption.

Conclusions

The magnetic link fault-current measuring device shows excellent correlation between its readings and corresponding values from oscillograms for both laboratory and field tests. The accuracy displayed on all the tests indicates a maximum deviation from the oscillographic record of approximately plus or minus 12 per cent and an average deviation of approximately plus or minus 5 per cent for the accepted design.

The tests conducted utilized a magnetic link and a surge-crest ammeter which were designed for a different purpose. Possibly a redesigned magnetic link and magnetism detecting device could be developed that would improve the operation of this device for the application to a-c measurements.

Magnetic Link Data. Midway Test May 20, 1945

Type of Fault	Number of Readings	Max Deviation,* Per Cent	Avg Deviation,* Per Cent
1ϕ-G	5	−10.5	2.7, −3.7
2ϕ-G	5	− 2.3	0.9, −1.5
ϕ-ϕ	6	6.6	3.3, −4.2

* Compare to asymmetrical rms current values.

The device may be used extensively on a system because of its simplicity and low cost.

Since it is a current-operated device, it may be used on lines or feeders of any voltage using the proper current transformers.

There are several advantages of the device, one being improvement over the accuracy of the annunciator ammeter for use as a fault locator using the current-ratio method. It may be used as a check on relay target operation when additional coils are used in the separate phases to determine the type of fault and phases involved. It is believed the device has very good capabilities for use on power systems for the determination of the magnitude of fault currents. It provides a method for determining the interruptoin current of an oil circuit breaker during a fault. The latter may be of value in scheduling maintenance for oil circuit breakers. The device measures the magnitude of the last half cycle of current being interrupted. The difference between the magnitude of the last half cycle cur-

Frequency Performance of Thyratrons

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Synopsis: The frequency performance of small thyratrons is investigated by supplying anode voltage from a variable frequency electronic generator of 1,400 watts output capacity. At high audio frequencies there is a departure from the 60-cycle-per-second performance; the grid control characteristic resolves itself into two characteristics—a starting characteristic and an extinguishing characteristic. The starting characteristic is shown to be a function of grid–anode capacitance and grid resistance. The extinguishing characteristic is determined by deionization effects and so is a function of frequency, grid resistance, anode current, tube geometry, and gas pressure.

THERE are many applications for thyratrons at frequencies above the power frequency of 60 cycles per second. Inverters, servo controls, relaxation oscillators, radar modulators, and grid-controlled rectifiers for airborne equipment are examples of such applications. The frequency range is from a few hundred to 50,000 cycles per second. Although the circuits used in these applications are well understood, the exact nature of ionization and deionization effects on thyratron performance generally has not been known. This paper reports an investigation of such effects as affected by anode-supply frequency. A sinusoidal wave shape was used for the various tests because it frequently is used in actual circuits and because it is easy to generate. Wave shape affects thyratron performance, but performance with other wave shapes can be extrapolated from the results obtained with sine waves.

In the tests the sinusoidal voltage is applied to the anode circuit of the tube under test; hence the voltage source must have considerable power capability. A high-power electronic audio-frequency generator capable of supplying 1,400 watts of power at 115 volts in the frequency range of 300 to 3,500 cycles per second was used in this investigation.[1] By means of an external signal generator, the equipment also could be used as a power amplifier to produce reduced power at frequencies up to 20,000 cycles per second. The output voltage of the generator is maintained at a preset level independent of load by means of a regulating circuit having a time constant of about one second.

Figure 1 shows a block diagram of the test setup. *T* is a stepup transformer to supply the anode voltage for the tube under test. R_L is a load resistor to limit the average anode current $\bar{I}_b$. Filament power for the tube under test comes from the 60-cycle power line. Peak anode voltage is measured with a diode type peak voltmeter. Wave forms of both current and voltage are monitored on a cathode-ray oscilloscope. Grid voltage is supplied by a 90-volt battery.

One type of thyratron, 3D22, was selected to be run through all the tests in which conditions would be subject to control by the user. The type 3D22 is a xenon-filled shield-grid thyratron rated for a peak forward voltage of 650 volts, a peak inverse voltage of 1,300 volts, and an average current of 0.75 ampere. Its control ratio is 150.

However, other tube types were put through some of the tests. The results reveal the effects of tube geometry and aid the user in selecting the proper tube type for a given application.

An additional test shows the effect of change in gas pressure, a characteristic which is subject to control by the user only in the case of mercury thyratrons.

Thyratron Control Phenomena

Transition to On-Off Control

At a low frequency (60 cycles per second), the average current through a thyratron can be controlled by the grid and can be calculated from the following equation,[2] which neglects tube drop:

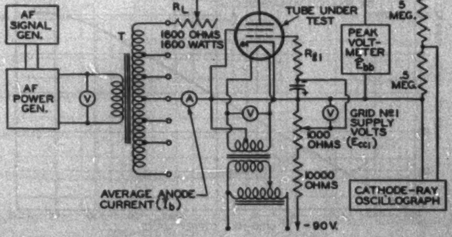
$$\bar{I}_b = \frac{\hat{E}_{bb}}{2\pi R_L}(1+\cos\theta)$$

$\hat{E}_{bb}$ = peak anode supply voltage (sinusoidal)
R_L = load resistance
θ = phase angle at which conduction begins

When the grid voltage is zero in a negative grid thyratron, θ is zero, conduction starts at the beginning of the anode voltage cycle, and maximum average current flows. With increasing negative grid voltage, θ approaches 90 degrees and the current is reduced to one-half the maximum average value. With only direct voltage on the grid, 0 to 90 degrees is the maximum range of variability since making the grid still more negative cuts off the current completely.

When the anode supply frequency is increased sufficiently, this mechanism breaks down. The grid no longer is able to vary the anode current; it becomes simply an on-off control, the current being established only by the external anode circuit. Figure 2 shows this transition in the case of the type 3D22 thyratron. The explanation of this form of operation is that ions remaining from the previous conduction cycle interfere with the grid control. At low frequency the interference is small and the extinction current is

Figure 1. Test setup for audio frequency testing of thyratrons



$\bar{I}$ = average current
$\hat{E}$ = peak voltage
Rating of transformer *T*:
Primary = 115 volts
Secondaries = 3,500/1,750 volts, 2,000/1,000 volts, and 1,000/500 volts
Volt-amperes = 2,120/2,750
Frequency = 0.3 to 3.5 kc

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rent and the current at the time the contacts part, which is defined as the interrupted current, is usually negligible. This may not be true in the case of very fast circuit breakers opening before the d-c component has decayed to a noncritical point, for restrikes, nor for low-voltage circuit breakers where the arc voltage is sufficient to choke down the last half cycle of current. The instances in which the device will be subjected to these conditions probably will be infrequent.

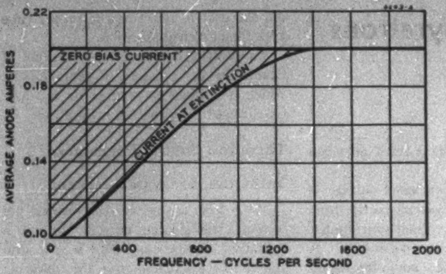

Figure 2. Extinguishing current versus frequency for type 3D22 thyratron

R_{g1} = 0.1 megohm
E_{cc2} = 0
R_{g2} = 0
$\hat{E}_{bb}$ = 650 volts

therefore nearly half the maximum current. As the frequency is increased, this interference increases, the extinction current increases until it reaches the maximum average value, and we have solely on-off control.

However, if one desires to obtain a variable anode current at frequencies beyond 1,500 cycles per second for the type 3D22 thyratron one can use the d-c bias-phase-control method.[3] In this case the grid voltage consists of a combination of direct voltage plus a 90 degrees lagging alternating voltage. This method is successful in controlling the average anode current at higher anode supply frequencies whereas the direct voltage method alone would fail. The explanation is that the alternating grid voltage sweeps out ions while swinging negative. Figure 3 gives the conditions under which this control may be applied to the type 3D22 thyratron at a frequency of 2,000 cycles per second where d-c grid voltage alone would not permit variable control.

TWO CONTROL CHARACTERISTICS—STARTING, EXTINGUISHING

It became apparent as these tests progressed that there is not one but two control characteristics. One characteristic describes the relation between anode voltage and grid voltage on starting.

Table I. Capacitance Measurements of the Type 3D22 Thyratron

Tube Component	Capacitance, μμf
Socket	0.229
Grid–anode (inside envelope)	0.094
Effective base pin (tube in socket)	0.126
Total	0.449

This is the initial conduction of current and is not to be confused with the every cycle starting which accompanies an a-c anode supply. The other describes this relation for extinction or cutoff of the discharge. By cutoff is meant failure to reignite on successive cycles. Figure 4 shows how the extinguishing characteristic has shifted considerably in the negative direction at 2,500 cycles per second. This shift is caused by a lack of complete deionization. At a low frequency (60 cycles per second) the extinguishing characteristic is coincident with the starting characteristic.

To the author's knowledge the first investigator to indicate that there are two control characteristics was W. G. Shepherd of the Bell Laboratories. In 1943 Shepherd published a paper describing work on deionization effects in thyratrons.[4] He differentiated between a static and dynamic characteristic. The work was done with alternating voltage on the grid, and the dynamic characteristic was in terms of peak a-c grid swing necessary for stable operation.

However, it is believed that the concept of extinction control characteristic as herein presented is closer to the accepted definition of control characteristic.[5]

THE EXTINGUISHING CONTROL CHARACTERISTIC

As would be expected the extinguishing characteristic is a function of frequency. Figure 5 shows how this characteristic shifts with frequency for the type 3D22 thyratron. At 3,500 cycles per second the grid cutoff voltage is nearly ten times the 60-cycle value. The shift is increasing at a rate faster than proportional to frequency. Nevertheless, the grid is still potent as a control electrode, and in none of these tests did the grid lose control completely. If the grid voltage is made sufficiently negative it will sweep out the ions at a rate such that it can prevent reignition on the next positive cycle. The limit is reached, however, when the sum of the negative grid voltage plus the peak anode voltage exceeds the cold breakdown value between grid and anode; in other words Paschen's law is operating.

THE STARTING CONTROL CHARACTERISTIC

The starting control characteristic for the type 3D22 thyratron was found to be constant for frequencies from 60 to 3,500 cycles per second. However, this condition is a special case; usually the starting control characteristic is a function of fre-

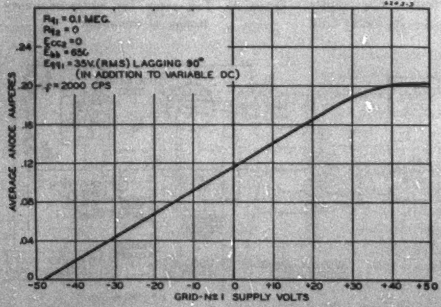

Figure 3. Thyratron control of type 3D22 combined direct voltage and 90-degree lagging alternating voltage

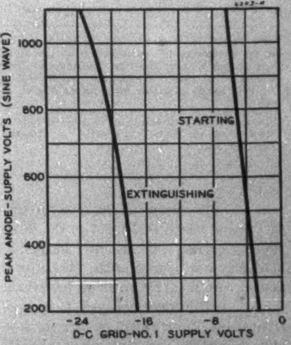

Figure 4. Control characteristics of type 3D22 thyratron starting and extinguishing

R_{g1} = 0.1 megohm R_{g2} = 0
I_b = 0.2 ampere at E_{cc1} = 0 E_{cc2} = 0
Frequency = 2,500 cycles per second

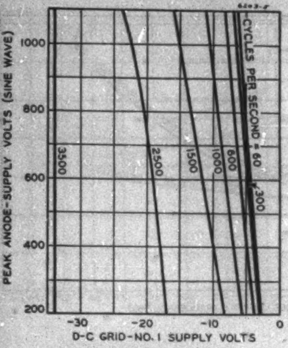

Figure 5. Typical extinguishing control characteristic of type 3D22 thyratron

$R_{g1}=0.1$ megohm $R_{g2}=0$
$I_b=0.2$ ampere at $E_{cc1}=0$ $E_{cc2}=0$

quency. In order to illustrate the general case consider the type 2050 thyratron. The type 2050 is a xenon-filled shield-grid thyratron similar to the type 3D22 but smaller in size. The ratings are a peak forward voltage of 650 volts, a peak inverse voltage of 1,300 volts, and an average current of 0.1 ampere. Its control ratio is 250. Figure 6 gives the starting control characteristic as a function of frequency for this type. One may observe that, like the extinguishing characteristic, the shift is toward the negative grid voltage but that the shift is not as great. This shift is not caused by ionization but by coupling to the grid through the anode–grid capacitance shown in Figure 7.

The anode–grid capacitance C_{pg} is small and therefore its reactance is large in comparison with the grid resistance. Hence the induced grid voltage is nearly 90 degrees ahead of the anode voltage. This induced voltage becomes tangent to the critical grid voltage at point *x*. As the frequency of the anode supply is increased the induced voltage increases, and thus conduction begins at a more negative d-c grid voltage.

Control Versus Grid Resistance

The tests presented thus far all were made at one value of grid resistance, namely 0.1 megohm. Figure 8 shows the effect of varying this grid resistance at a fixed frequency of 1,000 cycles per second. Both the starting and extinguishing grid voltages become more negative as the grid resistance is increased.

It is now clear why no shift in starting characteristic was observed in the case of the type 3D22 thyratron. The grid resistance chosen (0.1 megohm) was not sufficiently large.

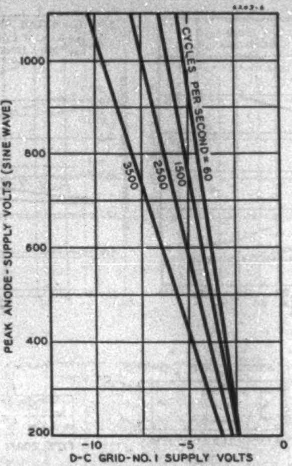

Figure 6. Typical starting control characteristic of type 2050 thyratron

$R_{g1}=0.1$ megohm $R_{g2}=0$ $E_{cc2}=0$

The explanation of the starting shift has been given. To check the data in Figure 8, capacitance measurements were made on the type 3D22 thyratron and its socket with the results as given in Table I. It is interesting to note that only one-fourth of the capacitance is contributed by the tube electrodes. Tube capacitance as published by tube manufacturers includes neither the interbasepin capacitance nor the socket capacitance.

The capacitive reactance of 0.449 micromicrofared at 1,000 cycles per second is 354 megohms. The peak induced grid voltage for the conditions of Figure 8 at five-megohm grid resistance is 9.18 volts. The induced grid voltage does not cut the critical grid voltage at the peak of the induced voltage, but it is close to this value. Reading the data from the curve of Figure 8, we find that the shift in control grid bias from 60 to 1,000 cycles per second is from minus 4.5 to minus 16.5 or 12 volts. The remainder of the shift is caused by direct current resulting from leakage or grid emission.

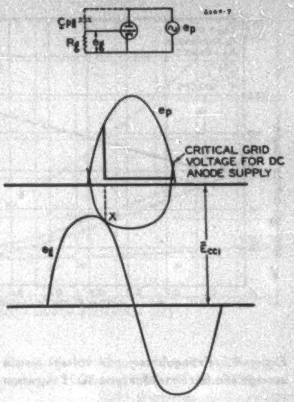

Figure 7. Anode–grid coupling in thyratron

The shift in extinguishing characteristic increases with grid resistance because of the effects just described plus deionization effects. Increasing the grid resistance decreases the effectiveness of the grid as a deionizing agent. The grid is able to sweep out fewer ions between conduction periods.

Control Versus Anode Current

It is a common conception that the control characteristic is independent of current. At high audio frequencies this is not true. The cutoff bias is a linear function of the average anode current as Figure 9 shows.

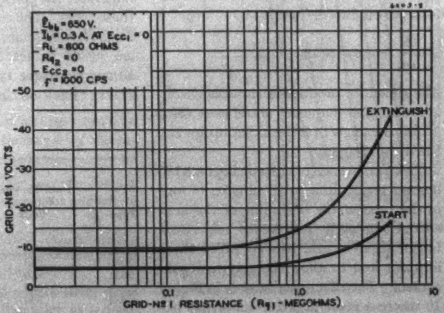

Figure 8. Critical grid bias versus grid resistance of type 3D22 thyratron

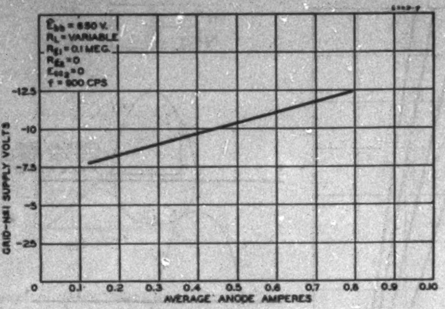

Figure 9. Extinguishing grid voltage versus average anode current for type 3D22 thyratron

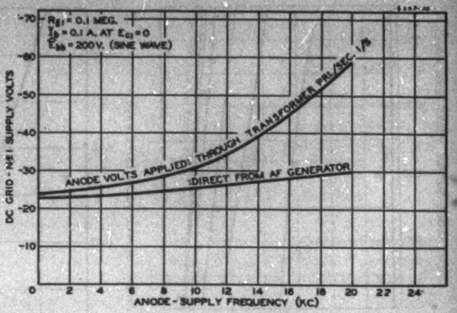

Figure 10. Effect of inductance in anode supply on extinction for type 884 thyratron

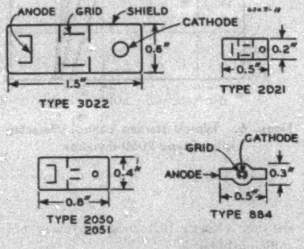

Figure 11. Thyratron constructions

Such a linear relationship would be expected if the following train of events are considered. The number of positive ions in the discharge is proportional to the current that has passed. Between conduction cycles the quantity of ions decays exponentially at a fixed rate, so that at a fixed time after current ceases in the external circuit the number of ions is proportional to the current that flowed. The negative grid, being the most effective deionization agent, sweeps the ions out at a rate approximately proportional to the negative applied voltage. At the beginning of the next cycle, if the grid has reduced the ions below a critical number, the anode rising in voltage will not be able to initiate a discharge.

Because of these relationships the anode current must be specified when the control characteristic at high frequency is given.

Effect of Inductance

When there is inductance in the load, the current persists after the anode voltage has dropped to zero. The solution[8] for a pure inductive load is

$$i_b = \frac{\hat{E}}{\omega L}(\cos\theta - \cos\omega t)$$

The generator produces a sinusoidal voltage of angular frequency ω and peak value $\hat{E}$. θ is the angle of firing as set up by E_{cc}, the d-c grid supply.

This persistence of current beyond 180 degrees results in a larger number of ions in the tube at the beginning of the next conduction cycle and hence requires a larger negative grid bias to sweep these ions out and to allow the grid to regain control. The effect may be seen in Figure 10 where a more negative curve was obtained when a transformer was inserted between the audio generator and the tube under test.

In taking these data the transformer ratio, resistive load, and voltage were selected to minimize the effect of transformer inductance.

Comparative Data on Different Thyratrons

Types 2051, 2D21, and 884 also were tested in this investigation in addition to types 3D22 and 2050. Ratings of these types appear in Table II. Figure 11 illustrates the construction of each one of these types. The cross sections are shown accurately to scale; the dimensions given are inside values. Note the similarity among the three types 3D22, 2050, and 2D21. Each of these types is a xenon-filled shield-grid thyratron. Types 2051 and 884 are argon-filled.

Starting Control Characteristic

In Figure 12 the starting control characteristics are compared for the four high-control-types. Types 2050 and 2051 have a high anode–grid capacitance and therefore exhibit a large shift. The 2D21 thyratron has the smallest grid–anode capacitance and hence the smallest shift. The frequency has to be increased to 15,000 cycles per second before a shift is observed in the 2D21.

Figure 14 gives the starting characteristic for type 884. This thyratron has a large grid–anode capacitance but its control ratio is so low that the shift is not excessive even at high audio frequencies.

Extinguishing Characteristic

The extinguishing control characteristic for the type 3D22 thyratron is shown in Figure 5. In Figure 13 the extinguishing characteristics are compared for the four high-control-ratio types. The shift in control is roughly proportional to the cross-sectional area of the type. Table III gives another indication that the phenomena are effects of deionization.

Although the type 2051 thyratron is a

Table II. Comparative Ratings of Thyratron Tubes

Type	Peak Forward Voltage, Volts	Peak Inverse Voltage, Volts	Average Current, Amperes	Anode–Grid Capacitance (Including Socket), μμf	Control Ratio
3D22	650	1,300	0.75	0.45	150
2050	650	1,300	0.100	0.60	250
2051	350	700	0.075	0.60	250
2D21	650	1,300	0.100	0.25	250
884	350	350	0.075	6.4	10

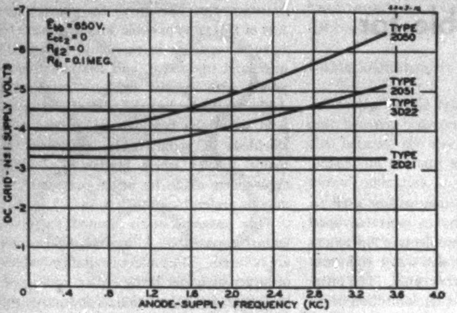

Figure 12. Comparison of starting grid voltages for types 2D21, 3D22, 2050, and 2051 thyratrons

renewal type, not to be specified for new equipment, it was tested because it was expected that the use of the lighter gas argon in place of xenon would lower its deionization time. However, the gas pressure is nearly three times greater in the latter tube and therefore counteracts the difference between gases.

Referring to Figure 14 for the type 884, it may be seen that the starting and extinguishing characteristics are identical. One must conclude that the effect of deionization on the control characteristic is negligible even up to 20,000 cycles per second. The whole shift is caused by grid-anode coupling; consequently, the type 884 is the most desirable tube for high frequency work.

Effect of Pressure

Starting Control Characteristics

A number of type 2050 tubes were made with pressures ranging from 25 to 250 microns of xenon. The starting characteristics of these tubes were measured, but there is no definite shift with pressure. Had the pressure range been extended down to about five microns a shift probably would have been observed since this is the pressure range in which mercury thyratrons exhibit a large shift in control characteristic.[6] The explanation for this shift in starting control characteristic of a thyratron with pressure will be considered in order to distinguish it from the other shift phenomena observed in this paper. It is the same as the explanation for Paschen's breakdown curve of a gas diode. As the pressure is increased each electron makes more collisions and produces more ions in crossing the tube. Thus the critical anode current for the formation of a discharge with the given anode voltage is reached at a lower initial electron current or at a greater negative grid voltage.

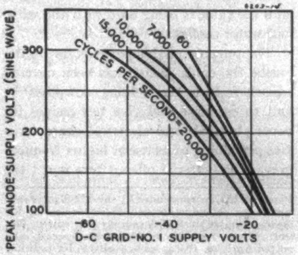

Figure 14. Starting and extinguishing characteristic of type 884 thyratron

R_{g1} = 0.1 megohm I_b = 0.1 ampere at E_{cc1} = 0

Extinguishing Control Characteristic

It is known that deionization time is a function of pressure[7] so that, if the shift in extinguishing control characteristic is a deionization effect, the shift should be a function of pressure. The results of the test for extinguishing characteristic are elegant as Figure 15 will attest. The curves fall in order toward the negative grid voltage as pressure is increased.

Conclusions

The following conclusions can be drawn about the performance of thyratrons at high audio frequencies:

1. The grid loses control over the average anode current but is still potent enough to shut off the discharge.

2. There are two control characteristics instead of one—a starting characteristic and an extinguishing characteristic.

3. The starting characteristic shifts negatively with increasing frequency as a result of coupling through grid-anode capacitance.

4. The extinguishing characteristic shifts negatively with increasing frequency because of coupling through grid-anode capacitance as well as because of deionization effects.

5. The extinguishing characteristic shifts negatively with increasing grid resistance.

6. The extinguishing characteristic shifts negatively with increasing anode current.

7. A comparison of tubes of similar construction indicates that the shift of extinguishing characteristic with frequency is roughly proportional to the cross-sectional areas of the tube cage.

8. The starting characteristic may shift

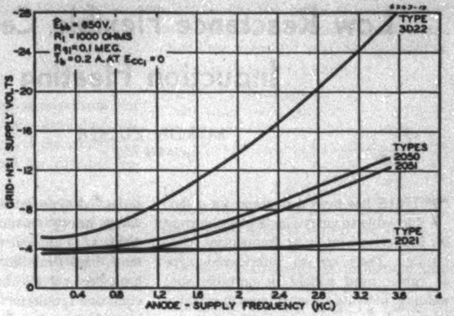

Figure 13. Comparison of extinguishing grid voltages for types 2D21, 3D22, 2050, and 2051 thyratrons

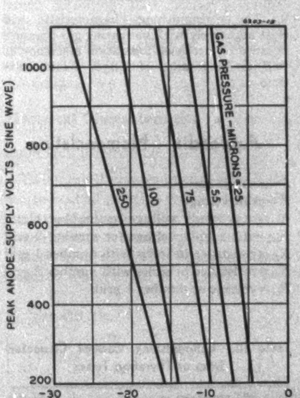

Figure 15. Extinguishing characteristic versus pressure for type 2050 thyratron

R_{g1} = 0.1 megohm R_{g2} = 0
I_b = 0.200 ampere at E_{cc1} = 0 E_{cc2} = 0
Frequency = 3,500 cycles per second

Low Reactance Flexible Cable for Induction Heating

MYRON ZUCKER
MEMBER AIEE

THERE has been a demand for a flexible cable to carry single phase current for induction heating at frequencies up to 10 kc. Tests on an interleaved type of cable, used widely in portable spot welding, indicate its usefulness in carrying thousands of amperes at potentials under 1,000 volts at the higher frequencies.

The flexibility of the cable has been sought for two purposes in induction heating:

1. For convenience of installation in the generator-transformer-coil circuit and for occasional shifting of the machinery.

2. For ability to carry currents (generally at low voltages) to coils that may be frequently or even continuously in motion.

In some cases, maneuverability is so important that it is advisable to use water cooled conductors in order to minimize cable size. Especially in such cases, it is logical to consider an adaptation of an interleaved type cable now widely used to carry heavy currents to portable spot welders. The basic problems of low reactance, high flexibility, and water tightness have been solved in this cable. The principal new requirement for induction heating is higher electric insulation. This has been attained for several hundred volts, and the cable is being used with and without water cooling.

The purpose of this paper is not to describe the cable (which has been covered elsewhere from the welding viewpoint[1,2]), but to estimate and give test values for some physical and electrical characteristics pertaining to its use at higher frequencies. The material offered here is not exhaustive, but introduces the characteristics of this type of cable which appears to have advantages in economy, low reactance and resistance, and flexibility compared with coaxial cable. Interleaved cable should not be used where its external field would be troublesome, unless it is shielded. Coaxial cable, described in many articles where studies of coaxial cables were made for other purposes,[3,4,5] may be preferable in such cases.

This presentation is limited to interleaved type cable. Physical dimensions are covered. They affect flexibility (cables of larger diameter being stiffer) and must be known for calculation of inductive reactance. The latter is covered in some detail since it is the major element of impedance of the cable at the frequencies involved. Some data are given on resistance, capacitance, and leakage losses. Cable ratings are not considered here since further information is needed on increased resistance losses at high frequencies in order to extend welding application data.[2] Also, more data are required on heating time constants at low water rates to evaluate heating and cooling of the cable under intermittent operations.[6]

Description of Cable

The essence of this cable design is that each conductor is broken up into a number of "ropes" (see Figure 1) placed around a central nonconducting core in a manner similar to the outer groups of the ordinary concentric-lay stranded electric cable. Ropes of alternate polarity lie next to each other except for a thickness of insulation between them. Since water cooling is desired each conductor is bare, and the insulation takes the form of a tube of irregular cross section that is formed in place inside half of the conductors and outside of the other half. The whole mass is held in a

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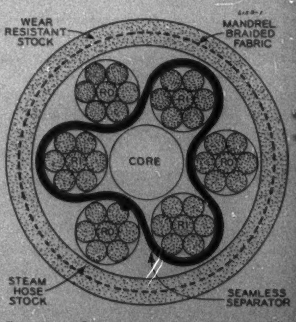

Figure 1. Cross section of interleaved cable for six ropes, three per electric side

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negatively with increasing gas pressure because of increased ionization efficiency.

9. The extinguishing characteristic will shift negatively with increasing gas pressure because of increased ionization efficiency as well as for reasons of longer deionization time.

Appendix. Nomenclature

I = average current
$\hat{E}$ = peak voltage
E_{cc1} = d-c supply voltage for number 1 grid
E_{cc2} = d-c supply voltage for number 2 grid
R_{g1} = resistance in series with number 1 grid
R_{g2} = resistance in series with number 2 grid
E_{c1} = voltage at number 1 grid
E_{c2} = voltage at number 2 grid
E_{gg1} = a-c (rms) supply voltage for number 1 grid
$\hat{E}_{bb}$ = peak supply voltage for anode
I_b = average anode current
f = frequency of anode supply

Note: All voltages are referred to cathode.

Table III. Extinguishing Control Characteristics of Thyratron Tubes

Type	Cross Section, Square Inches	Shift in D-C Supply Voltage for No. 1 Grid E_{cc1}, Volts
3D22	0.818	22
2050	0.284	9
2D21	0.118	2

60–3,500 cycles per second; $\hat{E}_{bb}$ = 600 volts.

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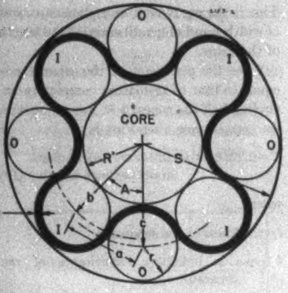

Figure 2. Symbols for analysis of cable section

regular pattern by means of an outer covering.

Some of the computations that follow are based on this arrangement. Then simpler formulas are derived from an approximate pattern in which all the conductors are on the circumference of one circle with radial insulating partitions. It is evident that the more ropes that are provided, the lower will be the reactance and the skin effect, which becomes important at higher frequencies. More ropes entail a larger core.

The core is a nonconductor which has been adopted for mechanical reasons even though the use of copper in the core would result in a smaller over-all cable diameter. The question is how great a price in terms of bulk must be paid for the lower impedance obtained by subdividing the copper.

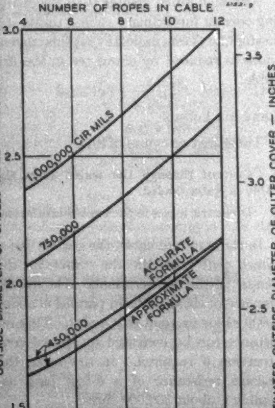

Figure 3. Calculated dimensions versus number of ropes in three sizes of interleaved cable

Cable Dimensions

The size of the cable is important not only because it determines flexibility, but also because the electrical characteristics depend on the size and the distance between conductors.

The geometry is somewhat complicated by the fact that the conductors are on the circumferences of two circles. The ropes O outside the insulating separator are on one circle; the ropes I that are inside are on another circle (see Figure 2).

Let

r = radius of each conducting rope

R' = radius of insulating core

t = thickness of insulating separator

Then let

a = straight line distance between centers of adjacent ropes $= 2r+t$

b = radius of circle upon which I ropes center $= R'+r$

c = radius of circle upon which O ropes center $= R'+r+t$

To solve the triangle abc, let

$$S = \frac{a+b+c}{2}$$

Then

$$S = \frac{2r+t+R'+r+R'+r+t}{2} = R'+2r+t$$

which, by coincidence, equals the radius of a circle around the cable.

If n equals the total number of ropes, then the angle between the centers of adjacent ropes is $A = 360°/n$.

Since

$$\sin\frac{A}{2} = \sqrt{\frac{(S-b)(S-c)}{bc}}$$

we obtain, by substituting for A, S, b, and c from the foregoing expressions,

$$\sin^2\frac{A}{2} = (\sin^2)\frac{180°}{n} = (S-b)(S-c)/bc = \frac{(r+t)r}{(R'+r+t)(R'+r)}$$

This may be solved for the diameter of the core:

$$2R' = -(2r+t) \pm \sqrt{(2r+t)^2 - 4r(r+t)\left(1 - \frac{1}{\sin^2\frac{180°}{n}}\right)} \qquad (1)$$

Then the diameter D of the composite, which is also the inside diameter of the outer covering (neglecting clearances required for assembly), may be expressed as follows in terms of r, t, and n:

$$D = 2S = 2R'+4r+2t = 2r+t+\sqrt{(2r+t)^2 - 4r(r+t)\left(1 - \frac{1}{\sin^2\frac{180°}{n}}\right)} \qquad (2)$$

The values of D for several sizes of cable (with 4 to 12 ropes) are shown in Figure 3.

A simpler approach is to assume that the centers of the ropes are on the circumference of one circle of radius R. The circumference of this circle is

$$2\pi R = n(2r+t)$$

whence

$$R = n\frac{(2r+t)^*}{2\pi} \qquad (3)$$

The outer diameter of the array is taken as the diameter of this circle plus the diameter of a rope and one thickness of the separator. Thus

$$D = 2R+2r+t = n\frac{(2r+t)}{\pi}+2r+t \qquad (4)$$

The closeness of this approximation when insulation thickness is small compared with the size of conductors is shown in Figure 3. These curves are based on theoretical values of r. In actual cases sizes may vary considerably because of different stranding and cabling practices. The curves are within reason, however, and illustrate the effect of varying the amount and subdivision of copper area in this design.

Electrical Characteristics

Inductive Reactance

The induced voltage in each rope is the line-to-neutral reactive drop. Using the method of counting flux linkages up to some distance u outside of the cable, the voltage induced in rope 1 per centimeter length is found to be

$$E_{L-N} = \omega i 10^{-8} \times \left[\frac{1}{20} + 0.2\, ln\left(\frac{u}{r}\cdot\frac{u}{a_3}\cdot\frac{u}{a_5}\cdots\frac{a_2}{u}\cdot\frac{a_4}{u}\cdot\frac{a_6}{u}\cdots\right)\right]$$

where a_n is the distance between the center of rope 1 and the center of the nth rope and i is the current per rope.

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* More accurately,

$$R = \sqrt[n-1]{\frac{a}{n}} = \sqrt[n-1]{\frac{2r+t}{n}}$$

Through the following eight simple transformations, we obtain the value of X_{LL}:

1. Change to common units of volts per foot (2.54×12 multiplier).
2. Line-to-line voltage $E_{LL}=2E_{LN}$.
3. Total current $I=ni/2$.
4. Use $\omega=2\pi f$.
5. Cancel the common u in the parentheses.
6. Convert from natural to common logarithms.
7. Coalesce 0.05 with the logarithmic term by use of conductor geometric mean radius $GMR=r/0.78$.
8. Transform to reactance: $X_{LL}=E_{LL}/I$.

Through these steps we obtain

$$X_{LL}=\frac{7.68f\times10^{-6}}{n}\times 0.46\log\frac{a_2a_4\ldots\ldots a_n}{0.78ra_3a_5\ldots\ldots a_{n-1}} \quad (5)$$

This cumbersome form was found (as shown in the appendix) to reduce to

$$X_{LL}=\frac{3.53f\times10^{-6}}{n}\times \log\frac{4R}{0.78nr}\text{ ohms per 1,000 feet} \quad (6)$$

This may be stated in terms of basic dimensions by substituting the value of R from equation 3. Then

$$X_{LL}=\frac{3.53f\cdot10^{-6}}{n}\log\frac{2(2r+t)}{0.78r} \quad (7)$$

or, since

$$r=\sqrt{\frac{2CM/n}{2{,}000}} \quad (8)$$

$$X_{LL}=\frac{3.53f\times10^{-6}}{n}\log 5.1\left(1+\frac{1{,}000t}{\sqrt{2CM/n}}\right)\text{ ohms line-to-line per 1,000 feet} \quad (9)$$

where

f = frequency in cycles per second

t = thickness of insulation between ropes

n = the total number of ropes in the cable

CM = circular mils of conductor in each electric side of the cable; that is, half the total copper in the cable

Equation 9 shows that the reactance is nearly inversely proportional to the number of ropes per conductor, and only slightly dependent upon conductor dimensions.

This relationship is illustrated in Figure 4 at 10 kc for one thickness of separator, a range of common cable sizes, and two to six ropes per side.

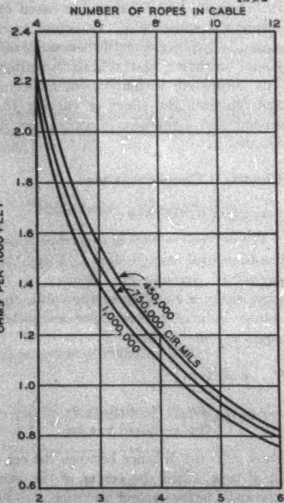

Figure 4. Inductive reactance at 10 kc versus number of ropes in three sizes of interleaved cable

RESISTANCE

No calculations have been made on skin effect or proximity effect of the array of conductors under discussion here. It is evident that the breaking up of the conductor and the interaction between ropes will produce a better distribution of current than in ordinary conductors, and so decrease the resistance.

Tests have been made on 6-rope cables at high power levels with direct current and at 60, 890, and 9,600 cycles. The latter were rather crude and are included only because they furnish an "order of magnitude" value. More accurate tests are planned at various frequencies and with different cable sizes and numbers of ropes.

From available tests, the approximate ratio of a-c resistance to d-c values are shown in Figure 5.

CAPACITANCE

The capacitance between the two electric sides of the cable depends upon whether it is dry or filled with water. An estimate is made most easily under the latter conditions, since the high dielectric constant of the water permits the first approximate assumption that the voltage stress is distributed uniformly across the rubber separator whose perimeter is p. This is the equivalent of a plate capacitor of width p and of length equal to the length of the cable.

Since the perimeter of the separator is equal to that of insulators completely encircling half the ropes, $p = 2\pi nr/2 = \pi nr$, the capacitance, therefore, is

$$C=0.224kp/t=0.224k\pi nr/t\text{ micromicrofarads per inch length} \quad (10)$$

where

t = thickness of the separator

k = specific inductive capacitance of the separator

Taking $k = 5$ for a rubber insulator, using a 1/16-inch thickness, changing to one foot of cable length, and substituting the value of r from equation 8, we get

$$C=0.34\sqrt{n}\sqrt{2CM}\text{ micromicrofarads per foot of cable length}$$

Considering this as a lumped capacitance,

$$X_c=\frac{1}{2\pi fc}=\frac{1}{2\pi f\times0.34\sqrt{n}\sqrt{2CM}}=\frac{0.47\times10^{6}}{f\sqrt{n}\sqrt{2CM}}\text{ megohm-feet of cable} \quad (11)$$

Figure 6 gives reactances calculated from this formula.

An idea of charging current is obtained by taking 500 volts across a 10-foot length of 6-rope 450,000-circular-mil cable at 10 kc:

$$I_c=\frac{500\times10}{20{,}000}=0.25\text{ ampere}$$

Although much greater than the charging current for ordinary cables, this is still negligible. Tests indicate that this current value is reduced by about ten in the dry cable.

LEAKAGE LOSSES

There are two causes of loss:

1. Current through the water when the cable is water cooled.
2. Dielectric losses in the rubber insulation.

In the standard cable, the current has a direct path through the water in the "caps" of the two terminals. The d-c resistance of these paths in parallel is about 6,000 ohms in good city water. This resistance can be increased by special construction if required. In dry cable the leakage resistance of a 6-foot long assembly is about 300,000 ohms.

The dielectric losses of insulation depend, of course, upon the material used. Neglecting losses in the water, a reasonably high value for present inferior insulation, is ten per cent of the charging current which has been derived previously:

$$\text{Dielectric loss} = \frac{0.10E^2}{X_c} \qquad (12)$$

For example, in the typical case of a 10-foot 6-rope 450,000-circular-mil cable, the dielectric loss at 500 volts, 10 kc will be about $0.10 \times 500 \times 500/20{,}000 = 12$ watts.

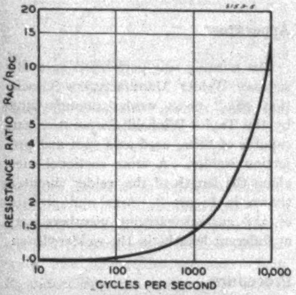

Figure 5. Approximate resistance ratio versus frequency for 6-rope interleaved cable

These losses are small enough so they can be neglected in most cases.

CONCLUSION

The electrical impedances of water cooled interleaved cables in the range of frequencies up to 10 kc are approximately

$$Z_{\text{series}} = \text{resistance ratio} \times \frac{23L}{CM} + j\frac{3.53f \times 10^{-9}L}{n} \times \log 5.1\left(1 + \frac{1{,}000t}{\sqrt{2CM/n}}\right) \text{ohms} \qquad (13)$$

$$Z_{\text{shunt}} = 6{,}000 - j\frac{0.47 \times 10^{13}}{fL\sqrt{n}\sqrt{2CM}} \text{ ohms} \qquad (14)$$

where

L = cable length in feet
CM = circular mils of conductor in each electrical side of cable
f = frequency in cycles per second
n = total number of ropes in the cable
t = thickness of separator in inches

As an example, when operating a 10-foot length of 6-rope 450,000-circular-mil cable at 10 kc,

$Z_{\text{series}} = 0.0064 + j0.0155$,
$Z_{\text{shunt}} = 6{,}000 - j2{,}000$

If carrying 2,000 amperes at 500 volts, voltage drop $= 12.8 + j\,31 = 33.6\underline{/168^\circ}$, and leakage current $= 0.1 + j\,0.2$

In this frequency range, leakage is unimportant, and estimates may be made on the basis of equation 13. This is borne out by installations at 900 cycles and at 10,000 cycles where production economy demanded a flexible conductor.

Any modification from the commercial 6-rope cables, that are considered for the purpose of decreasing the voltage drop, must be viewed in the light of decreased flexibility because of larger diameter, and increased cost as a result of more difficult and nonstandard manufacture.

Appendix

Another method of deriving equation 6 is through geometric means, that is,

$$L_i = 2\ ln\ \frac{G_{io}}{G_{ii}} \qquad (15)$$

where

L_i = inductance associated with the inner ropes
G_{io} = geometric mean distance between all the inner ropes and all the outer ones
G_{ii} = geometric mean distance of the inner ropes with respect to themselves

To state the G's in terms of cable dimensions the simplest approach, suggested by H. B. Dwight, is to apply Guye's theorem which states that for m points or circles spaced equally around a circular line of radius R, the geometric mean distance of one from the others is

$$g = Rm^{\frac{1}{m-1}} \qquad (16)$$

(In this development, g is used for the distance from one rope to others, G denoting the distance between a group of conductors and others.)

First consider the inner ropes, letting m equal the number of inner ropes. Then the geometric mean distance from one inner rope to the other inner ropes is

$$g_{io} = Rm^{\frac{1}{m-1}} \qquad (17)$$

To find the geometric mean distance of the complete array with respect to itself, we also refer to the well-known relation that for a circle

$$g_{is} = r\epsilon^{-1/4} = 0.78r = r' \qquad (18)$$

Then the geometric mean distance of the inner set of ropes is obtained from

$$G_{ii}^{m} = g_{is}g_{ii}^{m-1} \qquad (19)$$

$$= r'\left(Rm^{\frac{1}{m-1}}\right)^{m-1}$$

$$= mr'R^{m-1} \qquad (20)$$

Having found the "self" geometric mean distance G_{ii}, we proceed to the "mutual" geometric mean distance G_{io} between inner and outer wires. This is done indirectly (since Guye's theorem gives a direct solution for arrays of equidistant points like . . *iii*. or . . *ioio* . ., but not for . . *ooioo* . . .) by finding $g_{i(i+o)}$, the geometric mean distance from i to all other ropes, and then combining that with g_{ii} to obtain G_{io} from equation 22.

That is, using equation 16 with $2m$ = total number of ropes,

$$g_{i(i+o)} = R(2m)^{\frac{1}{2m-1}} \qquad (21)$$

Now because of the symmetry of the array, $g_{i(i+o)}$ also could be obtained by the process

$$g_{i(i+o)} = \sqrt[2m-1]{g_{ii}^{m-1}G_{io}^{m}} \qquad (22)$$

which gives

$$G_{io}^{m} = g_{i(i+o)}^{2m-1}/g_{ii}^{m-1} \qquad (23)$$

where G_{io} is the geometric mean distance of all the inner ropes with respect to all the outer ones. Substituting equations 17 and 21 in equation 23,

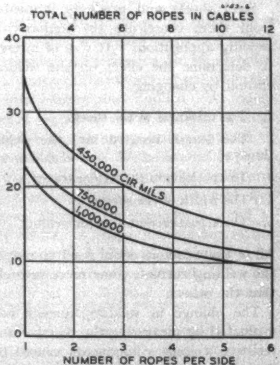

Figure 6. Calculated capacitive reactance at 10 kc versus number of ropes in three sizes of interleaved cable

$$G_{io}^{m} = R(2m)^{\frac{1}{2m-1}}/Rm^{\frac{1}{m-1}}$$
$$= 2mR^{2m-1}/mR^{m-1}$$
$$= 2R^{m} \qquad (24)$$

Combining equations 24 and 20,

$$\left(\frac{G_{io}}{G_{ii}}\right)^{m} = \frac{2R^{m}}{mr'R^{m-1}}$$

$$= \frac{2R}{mr'} \qquad (25)$$

$$m \cdot ln\frac{G_{io}}{G_{ii}} = ln\frac{2R}{mr'} \qquad (26)$$

Since $m = n/2$,

$$ln\frac{G_{io}}{G_{ii}} = \frac{2}{n}ln\frac{4R}{mr'} = \frac{2}{n}ln\frac{4R}{0.78mr} \qquad (27)$$

Influence of Magnetic Materials on the Welding Characteristics of Resistance Welding Machines

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MOST MEN associated with resistance welding machines are conscious of the fact that the introduction of magnetic material into the throat of such equipment causes a reduction of the welding current. The degree of reduction and simple mathematical tools for estimating it are contained in this paper.

Steel sheets and products fabricated from them constitute the majority of welding applications. It was of interest to determine the effect on the welding current by changing

1. The thickness of the sheets.
2. The length inserted into the welder throat.
3. The position in the welder throat.
4. The width of the sheets.
5. The actual current used for welding.

All of these factors could react to reduce the welding current, some more severely than the others.

The change in welding current was evaluated by measuring the reactive and resistive voltage components caused by the steel sheets in the welder throat. These voltage components in combination with the short-circuit impedance constants of the welding machine provide the necessary information.

Consequently, with a knowledge of these impedance constants and the nomogram indicating the effect of steel, it is a brief routine operation to calculate the change in welding current caused by the insertion of magnetic material into the secondary welder loop.

The apparent loss of current arises from the more intense magnetic field or increase in flux linkages and from the core loss resulting as heat generated in the steel sheets, a natural consequence of being subjected to an alternating magnetic field. Some effort was directed toward determining the magnitude and distribution of these flux linkages so that a better basic understanding of the influence of magnetic materials could be realized.

The magnitude of the magnetic field produced in the welder throat will depend primarily on the welding current. It is realized that a wide range of welding current can be used to make spot welds in mild or low carbon steel. The desirable maximum and minimum points of this range, together with other associated welding variables, are given in Table 1. Figure 1 summarizes the welding variables for seam welding of low carbon steel. Figure 2 lists the welding variables for spot welding of austenitic stainless steels.[1] The data shown in both Figures 1 and 2 are the tentative recommended practices of the American Welding Society.

These combinations of welding variables produce satisfactory spot welds and seam welds, and justify their use for further work of the influence of magnetic materials.

Apparatus

The welding was performed on a Resistance-Welder-Manufacturers-Association size-2 press welder manufactured by the Taylor-Winfield Corporation and capable of being used as a spot and projection welder. A cross sectional view along the length of the welder throat is shown in Figure 3. Area and position of the current carrying members vary at different lengths in the welder throat.

Procedure

It was desirable to make the necessary electrical measurements while making actual welds with the steel sheets inserted in the welder throat. However, the slight distortion of the current wave form caused by the volume of steel added to the circuit and the attempt to accurately read the small changes in power factor from a watt-galvanometer oscillographic trace made it imperative to separate the resistance caused by making a weld from the resistive component caused by the presence of steel sheets. It was evident that the welding resistance R_w was much larger than the resistive component R_{rs}, especially in thin sheets.

MEASUREMENT OF WELDING RESISTANCE R_w

The welding resistance R_w is the resistance from electrode to electrode across

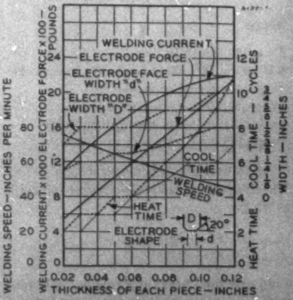

Figure 1. Low carbon steel seam welding data

Paper 46-137, recommended by the AIEE committee on electric welding for presentation at the AIEE summer convention, Detroit, Mich., June 24–28, 1946. Manuscript submitted March 11, 1946; made available for printing May 14, 1946.

J. J. RILEY and C. E. SMITH, a chief electrical engineer, are both of the Taylor-Winfield Corporation, Warren, Ohio.

Using equation 26 in equation 15, we find that

$$L_i = \frac{4}{n} \ln \frac{4R}{0.78nr} \text{ abhenries per centimeter} \quad (28)$$

Appropriate changes to practical units and conversion to reactance produce equation 6.

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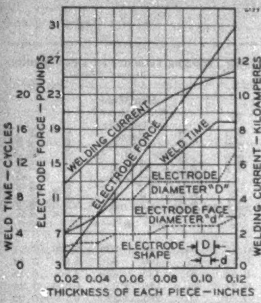


Figure 2. Austenitic stainless steel spot welding data

the work pieces when welding small coupons. It is the resistance added to the electrical circuit by welding operation itself and is independent of steel that extends into the welder secondary loop. R_w differs from the resistance R used in making the weld by that energy or heat that is dissipated into the work pieces and at the electrode faces.

Expressed mathematically

$$\text{Energy for welding} = I^2Rt + \text{losses}$$
$$= I^2R_wt$$

where

I = welding current
t = time of current flow
R = resistance used for welding
R_w = effective resistance present to the flow of welding current caused by welding

The value of R_w is the important quantity in welder operational characteristics, and it is fortunate that it can be measured easily while R is impossible to determine accurately.

The resistance R_w was measured when making a weld in small coupons by taking the voltage drop across the two electrodes at two points as close to the weld as possible and dividing this voltage by the secondary current. When welding stainless steel, the drop from the top of one sheet to the bottom of the other one (sheet to sheet resistance, V_2) also was measured. A schematic diagram of the setup is shown in Figure 4. The inductive pickup was minimized by carefully twisting the voltage leads together and holding them perpendicular to the electrodes.

Simultaneous oscillographic records were taken of the voltage drops, primary current, and watts. The primary current was converted into the secondary or welding current by multiplying by the turns ratio of the welding transformer. The magnetizing current was neglected being less than two per cent in the most exaggerated case.

The variations of R_w with time during spot welding for several thicknesses of low carbon steel with different combinations of welding variables is shown in Figures 5 and 6. A less comprehensive picture of the variation of R_w with time for austenitic stainless steel is shown in Figure 7. A similar set of data for seam welding low carbon steel is presented in Figure 8. By selecting average values of R_w for spot welding low carbon steel and stainless steel under the conditions specified by Table I and Figure 2, the variation of R_w with sheet thickness results as shown in Figure 9.

The mild steel used in this investigation was covered with a slight film of oil to prevent rusting, being previously hot rolled and pickled. Before welding the steel was wiped with a clean rag to remove any dirt that accumulated during storage. The surface was still perceptibly oily after wiping. This type of surface condition, most often used in production applications, increases R_w beyond that expected from degreased sheets.

Measurement of Resistive Voltage Component, V_{rs} and Reactive Voltage Component, V_{xs} Caused by the Insertion of Magnetic Material or Steel into the Welder Throat

These voltage increments were measured without involving the welding resistance R_w. A hole two inches in diameter with its center two inches from the front edge and midway across the width was drilled in two equal thicknesses of steel sheet (Figure 10). The sheets were insulated from each other by a sheet of paper, inserted in the welder throat midway between the welder horns, and locked in position with wooden blocks in order to counteract the magnetic force when the welder current was applied. A water-cooled copper tube was inserted through the hole in the sheets and connected to the welder horns. With the steel in place, the current necessary to make a weld and form an equivalent magnetic field was passed through the secondary circuit of the welder. A magnetic contactor, manually controlled, allowed interruption of the current whenever desired. The frequency of the applied voltage was 60 cycles per second.

Electrical values were measured using instruments equipped with pointer stops and capable of reading the rms values of distorted wave forms. Readings were taken of primary voltage, primary current, primary power, welder transformer turns ratio, and depth of steel in welder throat.

The steel sheets were prepared initially to extend the maximum depth into the welder throat. After each set of measurements six inches would be sheared off the

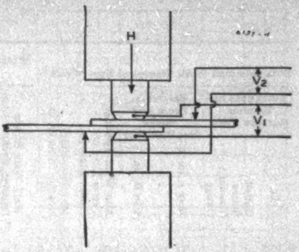


Figure 4. Electrode setup for measurement of welding resistance and sheet-to-sheet resistance

$$\text{Welding resistance } R_w = \frac{V_1}{I}$$

$$\text{Sheet resistance } R_{ss} = \frac{V_2}{I}$$

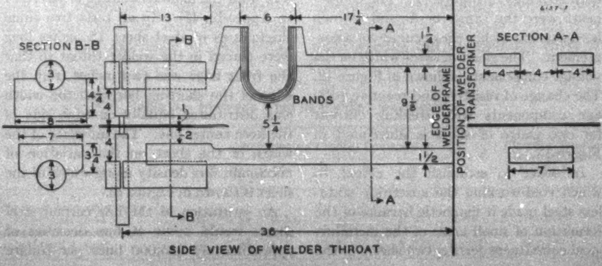


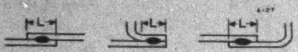
Figure 3. Side view of welder throat construction showing copper and copper-alloy current conducting members

Table I. Low Carbon Steel Spot Welding Data

Data Common to All Classes of Spot Welds						Welding Variables for Best Quality Class-A Weld					Welding Variables for Better Quality Class-B Weld					Welding Variables for Good Quality Class-C Weld				
Thickness of Thinner Outside Piece		Electrode Diam and Shape (See Figure 1)					Min Weld Strength and Electrode Force					Min Weld Strength and Electrode Force					Min Weld Strength and Electrode Force			
T, In.	United States Standard Gauge	Max *d*, In.	Min D, In.	Min Weld Spacing, Center Line to Center Line, In.	Min Contacting Overlap (Note 8), In.	Weld Time Single Impulse, Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.	Weld Time, (Note 8), Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.	Weld Time (Note 7), Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.
0.010	32	1/8	3/8	1/4	3/8	4	200	200	4,000	0.13	5	130	170	3,700	0.12	15	65	130	3,000	0.11
0.021	25	3/16	3/8	3/8	7/16	6	300	450	6,100	0.17	10	200	390	5,100	0.16	22	100	320	3,800	0.14
0.031	22	3/16	3/8	1/2	7/16	8	400	840	8,000	0.21	15	275	770	6,300	0.20	29	135	680	4,700	0.18
0.043	19	1/4	1/2	3/4	1/2	10	500	1,300	9,500	0.23	21	360	1,160	7,500	0.22	38	180	1,050	5,600	0.21
0.049	18	1/4	1/2	7/8	9/16	12	650	1,550	10,300	0.25	24	410	1,420	8,000	0.23	42	205	1,320	6,100	0.22
0.061	16	1/4	1/2	1 1/16	5/8	14	800	2,050	11,600	0.27	29	500	1,900	9,000	0.26	48	250	1,770	6,800	0.25
0.077	14	5/16	5/8	1 3/8	11/16	21	1,100	2,800	13,300	0.31	36	650	2,620	10,400	0.30	58	325	2,420	7,900	0.28
0.092	13	5/16	5/8	1 5/8	3/4	25	1,300	3,550	14,700	0.34	44	790	3,450	11,400	0.33	66	390	3,160	8,800	0.31
0.107	12	3/8	5/8	1 13/16	13/16	29	1,600	4,600	16,100	0.37	50	960	4,350	12,200	0.36	72	480	4,050	9,500	0.35
0.123	11	3/8	5/8	2	7/8	30	1,800	6,000	17,500	0.40	60	1,140	5,550	12,900	0.39	78	570	5,200	10,000	0.37

Notes:

1. Type of steel SAE 1010.
2. Material free from grease, scale, and dirt.
3. Welding conditions determined by thickness of thinnest outside piece *T*.
4. Data for total thickness of pileup not exceeding 4*T*.
5. Minimum spacing is that spacing for two pieces for which no special precautions are necessary to compensate for shunting of adjacent welds. For 3 pieces increase spacing 30 per cent.
6. Electrode material, class 2, minimum conductivity, 75 per cent copper; minimum hardness, 75 per cent rockwell *B*.
7. Weld times from 1 to 30 cycles consisted of a single impulse, from 32 to 60 cycles consisted of two equal impulses, and from 63 to 90 cycles consisted of three equal impulses.
8.



previous length until five inches extended into the welder throat.

The short-circuit constants or welder electrical constants without any steel in the welder throat but with the proper horn spacing were checked after each set of measurements. The resulting values of impedance, reactance, and resistance referred to the secondary side are designated as Z_0, X_0, and R_0, respectively.

The values of the voltage components caused by the addition of steel in the welder throat were determined in the conventional manner (Appendix I). The voltages V_{rs} and V_{xs} are the resistive and reactive components referred to secondary circuit.

These components are plotted for several thicknesses of sheet and different horn spacings in Figure 11. The components were the same for 8- and 16-inch horn spacing within the accuracy of measurement. The effect of sheet width on the voltage components is shown in Figure 12. The change of resistive and reactive voltage components with welding current for one gauge of sheet is illustrated in Figure 13.

In order to establish the extent to which cold-working the austenitic stainless steel made it magnetic because of the formation of small areas of the metallurgical constituent ferrite, two sheets of the largest thickness available were checked at various dimensions in the welder throat. Values of V_{rs} and V_{xs} are shown in Figure 14.

Measurement of Flux Patterns

While the absolute values of resistive and reactive voltage components caused by steel in the welder throat are the working tools in electrical characteristics calculations, a study of the changes in flux and flux pattern in the welder throat circuit is most beneficial to understanding the phenomenon and applying an engineering analysis to the problem.

The flux was measured by evaluating it from the voltage induced in a search coil. The method used for calculating flux from the induced voltage is described in Appendix III. In all tests two equal thicknesses of steel sheet 12 inches long were placed in the welder throat so that the front edge was two inches from the edge of the shorting bar, and the width was distributed equally on each side of the horn center line. The influence of the width of the sheet on the variation of maximum flux density along width of the sheet is shown in Figure 15.

As saturation of the low carbon steel sheets would occur at flux densities of approximately 125,000 lines per square inch, it is obvious that the normal welding current saturated the sheets at the center, or midway between the current carrying conductors. In order to determine how much the welding current could be decreased before the steel directly between the conductors would not saturate, two of the thickest sheets used in past work were subjected to different currents and the maximum flux density measured

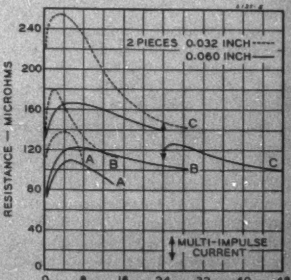


Figure 5. Variation of welding resistance with time in spot welding low carbon steel sheet

A, *B*, and *C* refer to classes of welds

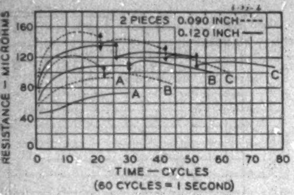


Figure 6. Variation of welding resistance with time in spot welding low carbon steel sheet

A, B, and *C* refer to classes of welds

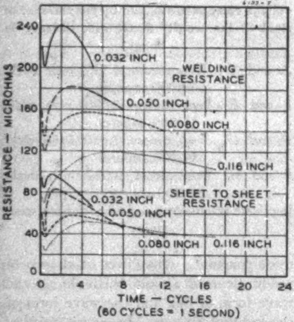


Figure 7. Variation of welding resistance and sheet-to-sheet resistance with time in spot welding austenitic stainless steel sheet of half-hard temper

(Figure 16). The thickest sheets were selected for this test as the ratio of welding current to sheet thickness decreases with increasing sheet thickness. Saturation still exists at the lowest value of current selected

Results

Welding Resistance R_w

The welding resistance R_w consists of two main components

1. The resistance from electrode to sheet.
2. The resistance from sheet to sheet.

The first component consists primarily of contact resistances. The second component consists of a contact resistance between the sheets and the resistance of the sheets themselves. From the test results on spot welding stainless steel in which these components were measured individually, it is notable that the resistance from sheet to sheet is approximately only 40 per cent of the total. It is expected that a similar relationship also would prevail for mild steel. Some past investigations[2-4] have indicated this approximate relationship. It is evident that the contact resistance between electrode and work piece is an appreciable part of the welding resistance and causes considerable heat generation at the electrode faces. Such being the case, it is very essential to cool the electrodes adequately in spot welding to prevent excessive deformation and to have the data apply.

While it is instructive to determine the components of the welding resistance R_w individually, the basic effect on the electric circuit is summarized in the value of R_w. From past experience it would be anticipated that R_w would be a function of the current density at the electrode face and the force density or pressure at the electrode face. As the variation of R_w with time could not be related satisfactorily to these two factors, average values of R_w were selected from Figures 5 and 6. Using these values and the associated data contained in Table IV, the following empirical relationship was determined for spot welding low carbon steel:

$$R_w = 333\,\frac{(D^{15}+1.7)}{P+2}$$

where

R_w = welding resistance in microhms
D = current density in (amperes per square inch) $\times 10^{-3}$
P = pressure in (pounds per square inch) $\times 10^{-3}$

The agreement between the calculated and measured values of R_w also is shown in Table IV. The equation gives passable results and point out that the heat generated in welding increases more rapidly than the square of the current, and reduction in electrode force causes appreciable increase in welding resistance.

In the empirical equation for R_w, the resistivity of the steel is not shown explicitly. However, it is erroneous to assume that the resistivity has no effect on welding resistance. In spot welding stainless steel, the values of R_w are considerably larger than those for mild steel of like gauges. The difference in resistivity, 12.3 and 73 microhms per centimeter cube, and surface condition between commercial hot rolled pickled and oiled mild steel and bright finished stainless steel, account for the marked differences in welding resistance between these materials.

The values of welding resistance in seam welding should depend on the same factors as in spot welding but the different nature of the electrode contact area while welding and the shunting effect of

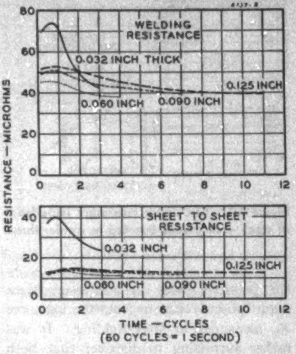


Figure 8. Variation of welding resistance with time in seam-welding low-carbon steel sheet

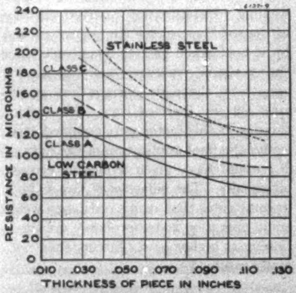


Figure 9. Variation of average welding resistance with thickness of sheet in spot welding

Table II. Materials Used in Welding Investigation

Type of Steel	Approximate Chemical Composition C	Si	Mn	Cr	Ni	P	S	Ultimate Tensile Strength, Lbs Per Sq In.
Commercial-quality cold-rolled steel sheet	0.08	0.34	0.35			0.01	0.01	45,000
Austenitic stainless steel, half-hard temper	0.10	0.50	1.1	17.3	7.3	0.015	0.006	160,000

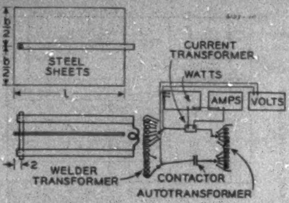


Figure 10. Circuit for measuring impedance of steel work pieces inserted in welder throat

the previous welds in a pressure tight seam cause deviations from the values of R_w measured in spot welding. It was rather surprising to discover that both the welding resistance R_w and the sheet to sheet resistance R_s, were almost equal in the range of thicknesses from 0.060 to 0.120 inch.

The emphasis on the determination of welding resistance was justified entirely as the resistive voltage component caused by it is of much greater magnitude than the resistive voltage component V_{rs} caused by the eddy current loss in the steel sheets inserted into the welder throat. If this factor were assumed carelessly, the final application of the data of V_{rs} would have given avoidable erroneous results. The values of R_w are so high in some cases that many resistance welders in operation today probably are functioning at higher power factors than normally assumed.

THE REACTIVE VOLTAGE COMPONENT V_{xs} AND THE RESISTIVE VOLTAGE COMPONENT V_{rs} CAUSED BY THE PRESENCE OF STEEL IN THE WELDER THROAT

Engineers making electric circuit calculations are familiar with the procedure of considering a resistance welder secondary loop in terms of reactive and resistive impedance constants. These are combined with impedance values evaluating the welding effect and the totals placed as the loaded secondary circuit of the welder transformer. This procedure and much factual data on welder secondary circuits are contained in an article by J. H. Cooper.[8] This practice must be altered when evaluating the presence of steel in the welder throat as ohmic values of reactance and resistance. These impedances change with the welding current. The effect of steel in the throat is to produce reactive and resistive voltage components opposing the applied secondary voltage of the transformer. These voltage components are almost independent of current over the normal welding range (Figure 14). Thus

Table III. Impedance of Welding Machine and Welding Transformer at a Frequency of 60 Cycles Per Second

Horn Spacing, Inches	Throat Depth, Inches	Impedance Z_o, Microhms	Reactance X_o, Microhms	Resistance R_o, Microhms
8	36	273	268	54
11	36	344	340	56

Note: There is no measurable change in impedance constants at secondary currents of 10,000 and 20,000 amperes.

$$V_{xs}=IX_s=C_1 \quad V_{rs}=IR_s=C_2$$

Consequently, if ohmic values of reactance and resistance are desired to express the presence of steel in the welder throat, the values must be selected in combination with the welding current to give the correct voltage drop.

Saturation of the steel sheets by the welding current is the prime reason why V_{xs} and V_{rs} remain almost independent of current. An elementary study of the flux changes in the steel sheets is instructive in pointing out the limitations of the data.

The magnetic field in air between two parallel conductors is shown in Figure 17. This field will be distorted when steel sheets are added, but the flux still will concentrate at the section of steel directly between the conductors. It must be realized that the effect of the steel sheets in distorting the air flux field as well as the effect of causing more flux linkages have been combined to evaluate V_{xs} and V_{rs}. In other words, the values of welder shortcircuit impedance constants X_o and R_o have been assumed constant in spite of the air field distortion and the higher harmonics in the welding current caused by the presence of steel sheets in the welder throat. This assumption is valid providing V_{xs} and V_{rs} do not become too large a part of the applied secondary voltage as will be the case in most welder secondary circuit construction.

The steel sheets are saturated by the magnetizing force caused by the welding current. The relationship of welding current, flux, induced voltage, and the flux-current curve for a specific thickness and length of steel are shown diagrammatically in Figure 18. As the welding current increases, the flux hardly increases as the intrinsic flux density B_i ($B-\mu_o H=B_i$) has reached its limiting value.* The area of the induced voltage curve is a measure of the maximum flux, and as the maximum flux under saturated conditions will be almost constant the average value of induced voltage will be almost constant. The effect of change of welding current should be to change the wave form of flux and the wave form of the induced voltage. In circuit calculations it is the effective value and not the average value of induced voltage that must be investigated. The measurements indicated that both the effective and average values of induced voltage stayed almost constant, in spite of change of welding current. In order to have this condition prevail with such wave forms of induced voltage as recorded, it is necessary to have the flux current relationship change with different magnetizing forces.

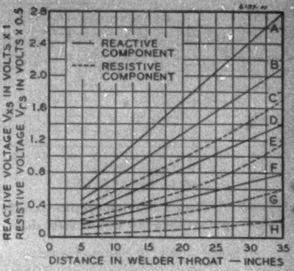


Figure 11. Variation of reactive and resistive voltage components caused by the presence of steel in the welder throat with distance in throat

A and C. Two pieces 0.119 inch thick, 18,000 amperes

B and E. Two pieces 0.090 inch thick, 15,000 amperes

D and G. Two pieces 0.060 inch thick, 12,000 amperes

F and H. Two pieces 0.032 inch thick, 8,300 amperes

Duplicate results for 8-and 16-inch horn spacings

The value of V_{xs} is proportional to the increase in length of steel in the welder throat. Under saturated conditions the maximum flux density is constant, and the increase in induced voltage is the result of the increase in flux caused by the increase in cross-sectional area provided by the additional length for a specific thickness.

Horn spacing or the distance between the throat bars does not influence V_{xs} appreciably. The steel sheets become saturated at any specific position in almost the same length of time from the zero point of the current wave. There is no

---

* The intrinsic flux density B_i is a measure of that part of the flux density attributable to the ferromagnetic characteristic of the material. When the material becomes saturated, it can make no further contribution to the flux density. As the magnetizing force H is increased beyond the value which saturates the material, the slope of the $B-H$ curve becomes μ_o, the permeability of free space, and B_i becomes constant.

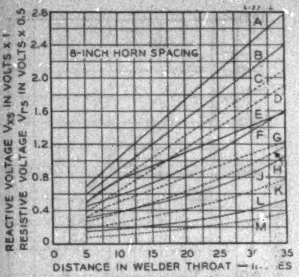


Figure 12. Variation of reactive and resistive voltage components with distance in throat and width of sheet

A, B, and *E.* Reactive voltages for two pieces 0.118 inch thick and widths of 36, 18, and 9 inches, respectively

F, J, and *L.* Resistive voltages under the above conditions

C, D, and *G.* Reactive voltages for two pieces 0.090 inch thick and widths of 36, 18, and 9 inches, respectively

H, K, and *M.* Resistive voltages under the same conditions

need to have the sheets equally spaced between welder horns to have the data apply.

It might be anticipated that V_{xs} would be independent of the width of the steel sheet if the steel were saturated at the section under the center line of the welder horns and thus limited the flux. By considering the air flux pattern, it is clear that narrow sheets will coerce less flux lines to enter the sheet than wide ones. As the current and magnetizing force change with time in the normal periodic manner, the flux will follow a similar pattern. As the saturation density is reached the preference for flux to pass into the sheet is eliminated. However, with narrow sheets it takes a longer period of time of a half-cycle of current flow to reach the saturation density. The resulting induced voltage is smaller as the rate of change of flux with time is smaller. It would be anticipated that after a certain width of sheet is reached, greater widths would cause negligible increase. This factor is associated with the distance between throat conductors or horn spacing and shape of conductors. The correction factors in Figure 20 are based on 8-inch horn spacing.

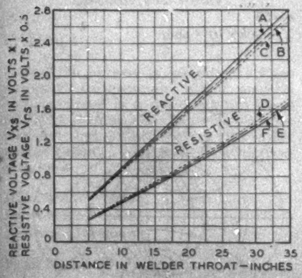


Figure 13. Variation of reactive and resistive voltage components with distance in throat and welding current. Two pieces of low carbon steel 0.118 inch thick and 36 inches wide

A and *F.* 19,000 amperes
B and *E.* 15,000 amperes
C and *D.* 10,000 amperes

Austenitic stainless steel is considered nonmagnetic, but cold working the steel causes the formation of small areas of iron or ferrite. These areas of ferrite would certainly reach saturation under the magnetizing force caused by the welding current. However, for the half-hard condition, the resulting values of V_{xs} and V_{rs} are small. The full-hard condition would have raised V_{xs} and V_{rs}. A word of caution is advisable in considering the magnetic effect in stainless steel. Those which are austenitic can be considered nonmagnetic for all practical purposes. However, those which are ferritic will be magnetic and should be considered having effects similar to mild steel.

The resistive voltage component multiplied by the secondary current represents the real power or core loss in the steel sheets. The great thickness of the sheet implies that the principle loss will be in the form of eddy currents. In the measurements, one steel sheet was insulated from the other, which means that the eddy currents would reach their maximum value near the outer surfaces of the sheets. If the sheets were in such intimate contact that the contact resistance between them could be considered zero, the resistive voltage component would increase theoretically to four times its former value. In most welding conditions the first case is more likely to exist than the second one, although some discretion should be used in applying the data.

The resistive voltage component caused by the presence of steel in the welder throat is normally a small part of the total resistive component composed also of the welder component I_tR_o and the component finished by the welding resistance I_tR_w.

Application of the Data

After making an analysis of the data and the limitations imposed upon them, it was desirable to compile these data into usable form. The values of V_{xs} and V_{rs} in Figure 11 were examined. By drawing straight lines for the V_{rs} variation with length, the empirical formulas and the resulting nomogram shown in Figure 20 were derived. The value of V_{rs} does not include the effect of welding resistance. The correction for width was obtained from Figure 12. The correction for difference in welding current from the base current was obtained from Figure 13.

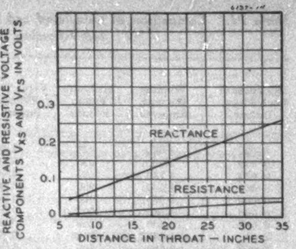


Figure 14. Variation of reactive and resistive voltage components with distance in throat for two pieces of austenitic stainless steel 0.109 inch thick, 24 inches wide and half-hard temper

The nomogram can be used for either spot welding or seam welding. For this reason, the effect of welding resistance is not included. The value of resistive voltage caused by the welding resistance can be selected from Figure 19. It is advisable to compare the desired welding technique with the technique outlined in Table I and Figures 1 and 2 so that a satisfactory choice of welding resistance voltage can be made.

Two examples of application of the data are shown in Appendix II.

Conclusions

1. The steel plates inserted in a welder throat will be saturated by the magnetic field produced by the welding current under normal conditions of operation.
2. The effect of the presence of steel plates (magnetic material) in the welder throat is measured best by expressing it as a reactive and a resistive voltage component. These voltage components are almost independent of welding current over the normal spot welding range.
3. The normal procedure of expressing the electric circuit of a welder in terms of resistance and reactance should be altered when the effect of steel plates is included. The circuit should be expressed in terms of reactive and resistive voltage drops, only some of which will be proportional to the current.
4. The effect of welding resistance and its

Table IV. Comparison of Calculated and Measured Values of Welding Resistance and Factors Affecting Welding Resistance

Thickness of Each Piece, Inches	Welding Current, Amperes	Electrode Force, Pounds	Force Density, Pounds Per Square Inch	Current Density, Kiloamperes Per Square Inch	Welding Resistance R_w, Microhms	
					Measured	Calculated
0.032	8,000	425	15,400	291	125	128
0.032	6,400	280	10,300	233	150	146
0.032	4,730	135	4,900	172	190	189
0.060	11,500	800	16,300	234	102	97
0.060	9,000	490	10,000	183	115	116
0.060	6,850	245	5,000	139	140	158
0.090	14,300	1,250	16,300	187	78	78
0.090	11,400	770	10,000	149	102	98
0.090	8,520	380	5,000	112	132	137
0.120	17,120	1,800	16,300	155	69	66
0.120	12,800	1,110	10,000	116	90	82
0.120	10,000	550	5,000	91	126	122

resulting voltage component is normally larger than the resistive voltage component caused by the presence of steel sheets for gauges less than 0.125 inch. As the thickness increases the effect of welding resistance voltage drop decreases and the resistive effect of the sheets becomes the predominating factor.

5. The effect of magnetic material in the welder throat is a function of the thickness and width of the sheets and of the secondary circuit constants of the welder. Resistance welders with large horn spacing are less influenced than welders with small horn spacing for the same throat depth. In other words, the larger the secondary voltage to supply a necessary welding current, the less the influence of the constant bucking voltage caused by the steel sheets.

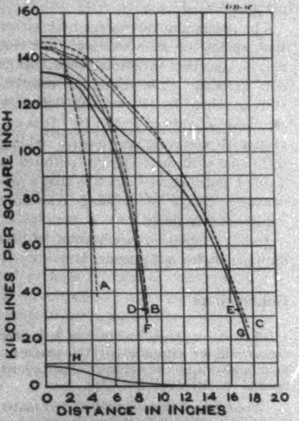


Figure 15. Variation of maximum flux density with distance from center line of electrodes at right angles to the welder throat

A, B, and *C.* Sheets 0.120 inch thick and 9, 18, and 36 inches wide, respectively, at 18,000 amperes

D and *E.* Sheets 0.090 inch thick and 18 and 36 inches wide, respectively, at 15,000 amperes

F and *G.* Sheets 0.032 inch thick and 18 and 36 inches wide, respectively, at 8,300 amperes

H. For magnetic field in air at 17,500 amperes

Appendix I. Determination of V_{rs} and V_{xs} from Instrument Readings

V_p = primary voltage, volts
I_p = primary current, amperes
W_p = primary power, watts
l = depth of steel in throat, inches
a = transformer turns ratio
V_t = secondary voltage, volts
I_t = secondary current, amperes
W_t = secondary power, watts
Z_p = impedance referred to primary winding, ohms
R_p = effective resistance referred to primary winding, ohms
X_p = reactance referred to primary winding, ohms
R_t = impedance referred to secondary winding, ohms
R_t = effective resistance referred to secondary winding, ohms
X_t = reactance referred to secondary winding, ohms
Z_s = impedance (for a specific current) of steel sheets referred to secondary winding, ohms
X_s = reactance (for a specific current) of steel sheets referred to secondary winding, ohms
R_s = resistance (for a specific current) of steel sheets referred to secondary winding, ohms
V_{rs} = resistive voltage component caused by steel sheets, volts
V_{xs} = reactive voltage component caused by steel sheets, volts
Z_0 = impedance of welder secondary circuit under short-circuit conditions, ohms
X_0 = reactance of welder secondary circuit under short-circuit conditions, ohms
R_0 = resistance of welder secondary circuit under short-circuit conditions, ohms

$$Z_p = V_p/I_p = a^2(Z_0+Z_s) \tag{1}$$
$$Z_s = (Z_p/a^2) - Z_0 \tag{2}$$

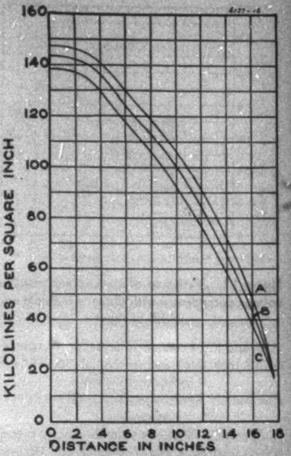


Figure 16. Variation of maximum flux density with distance from center line of electrodes at right angles to the welder throat and with welding current for two pieces 0.120 inch thick, 36 inches wide

A. 18,500 secondary amperes
B. 14,400 secondary amperes
C. 10,000 secondary amperes

$$R_p = W_p/I_p^2 = a^2(R_0+R_s) \tag{3}$$
$$R_s = (R_p/a^2) - R_0 \tag{4}$$
$$X_p = \sqrt{Z_p^2 - R_p^2} = a^2(X_0+X_s) \tag{5}$$
$$X_s = X_p/a^2 - X_0 \tag{6}$$
$$I_t = I_p(a) \tag{7}$$
$$V_{rs} = I_tR_s \tag{8}$$
$$X_{xs} = I_tX_s \tag{9}$$

The foregoing equations are used for the calculations of two sheets 0.118 inch thick, 18 inches wide, and inserted 35 inches in the welder throat.

$a = 52$
$X_0 = 268 \times 10^{-6}$ ohm

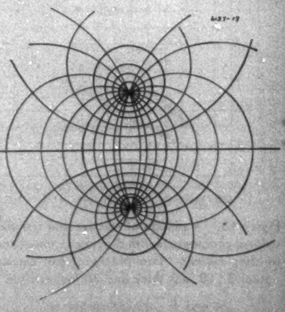


Figure 17. Graphical flow map of magnetic field in air between two parallel conductors

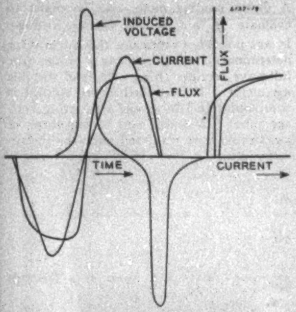


Figure 18. Schematic diagram of induced voltage, welding current, and flux in a saturated steel sheet

$R_o = 54\times10^{-6}$ ohm
$Z_o = 273\times10^{-6}$ ohm
$Z_p = 375/349$
$= 1.075$ ohms
$Z_s = 1.075/(52)^2 - 273\times10^{-6}$
$= 128\times10^{-6}$ ohm
$R_p = 26{,}700/(349)^2$
$= 0.219$ ohm
$R_s = 0.219/(52)^2 - 54\times10^{-6}$
$= 27\times10^{-6}$ ohm
$X_p = \sqrt{(1.075)^2 - (0.219)^2}$
$= 1.05$ ohms
$X_s = 1.05/(52)^2 - 268\times10^{-6}$
$= 121\times10^{-6}$ ohm
$I_t = 349(52)$
$= 18{,}100$ amperes
$V_{rs} = 18{,}100(27\times10^{-6})$
$= 0.488$ volt
$V_{xs} = 18{,}100(121\times10^{-6})$
$= 2.19$ volts

Appendix II. Applying Data to Specific Problems

Problem 1

Find the increase in secondary voltage necessary to maintain 18,000 amperes welding current in spot welding two sheets 0.120 inch thick and 36 inches wide when one inch and then 36 inches is inserted in the welder throat. Welder short-circuit constants are

$Z_o = 273\times10^{-6}$ ohm
$X_o = 268\times10^{-6}$ ohm
$R_o = 54\times10^{-6}$ ohm

From Figure 20

$I_t R_w = 1.26$ volts

From Figure 21

$V_{rs} = 0.86$ volt
Welder resistance drop $= I_t R_o$
$= (18{,}000)(54\times10^{-6})$
$= 0.97$ volt

Total resistive voltage drop $= V_r$
$= 3.09$ volts
Welder reactance drop $= I_t X_o$
$= (18{,}000)(268\times10^{-6})$
$= 4.82$ volts

From Figure 21

$V_{xs} = 2.87$ volts

Total reactive voltage drop $V_x = 7.69$ volts
Adding vectorially, $V_r + V_x = V_t$

$V_t = \sqrt{V_r^2 + V_x^2}$
$= 8.3$ volts at 36 inches
Voltage at start (one inch in throat) in which V_{rs} and V_{xs} are not effective
$= 5.32$ volts
Increase in voltage $= 8.3 - 5.3$
$= 3.0$ volts

Problem 2

The question is the same as in problem 1 except that the sheets are 18 inches wide, welding current is 10,000 amperes. The sheets are welded in the same welder. From Figure 20

$I_t R_w = 1.23$ volts

From Figure 21

$V_{rs} = 0.86$ volt

From this same chart this value must be corrected to

$K_1 \cdot V_{rs} = 0.65\ (0.86)$
$= 0.559$

because the sheet is only 18 inches wide. This value should be corrected further for change in base current, $K_3 = 1.05$.

$K_1 \cdot K_3 \cdot V_{rs} = (0.65)(1.05)(0.86)$
$= 0.587$ volt
Welder resistance drop $= I_t R_o$
$= 10{,}000(54\times10^{-6})$
$= 0.54$ volt
Total resistive voltage drop $= V_r$
$= 2.36$ volts

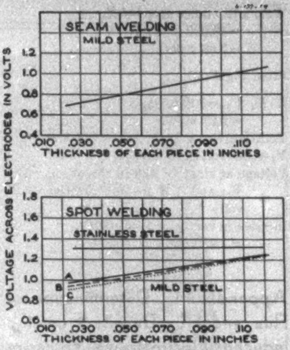


Figure 19. Resistive voltage drop across the electrodes caused by the welding resistance when welding small coupons

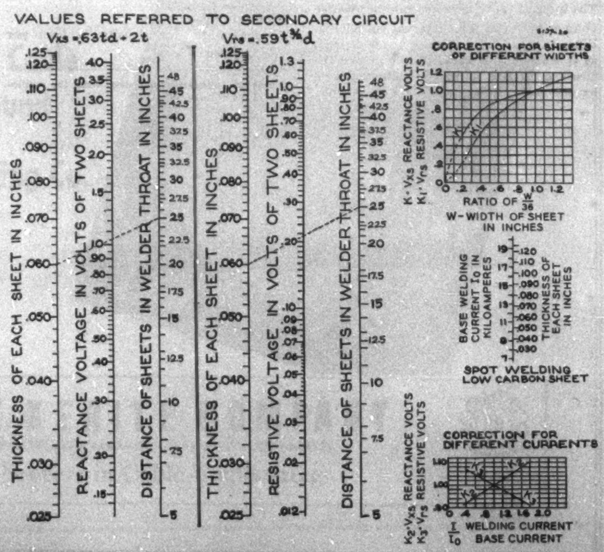


Figure 20. Reactive and resistive voltage components caused by two thicknesses of low carbon steel 36 inches wide in welder throat

Welder reactance drop $=I_t X_s$
$=10{,}000\,(268\times10^{-6})$
$=2.68$ volts

From Figure 21

$K\cdot K_2\cdot V_{xs}=0.84(0.94)(2.87)$
$=2.27$ volts
Total reactive voltage drop $=4.95$ volts
$V_t=5.48$ volts at 36 inches
Voltage at start (1 inch in throat) in which V_{rs} and V_{xs} are not effective $=3.21$ volts
Increase in voltage $=5.48-3.21$
$=2.27$ volts

Appendix III. Determination of Flux and Flux Density From the Induced Electromotive Force or Voltage Wave

Basically

$$e=-N\frac{d\phi}{dt}10^{-8}$$

$$d\phi=\frac{-1}{N}e\cdot dt\,\frac{1}{10^{-8}}$$

$$\phi=\frac{-10^8}{N}\cdot\int_{t_1}^{t_2}e\cdot dt$$

where

N = number of turns on search coil
e = instantaneous voltage, volts
$\frac{d\phi}{dt}$ = rate of change of flux with time
ϕ = flux, maxwells or lines
t = time, seconds

The integral $\int_{t_1}^{t_2}e.dt$ represents the average voltage over the time interval considered. When the maximum value of flux is desired, as it was in this work, it is necessary to measure the area under the voltage wave e for one-half cycle by means of a planimeter, divide it by two, and substitute in the expression

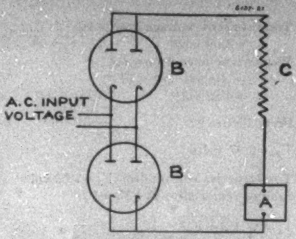


Figure 21. Rectifier unit for converting a-c input voltage to average direct voltage

A. D-c millivoltmeter, 1,000 millivolts, full scale, 5,000 ohms resistance
B. 6H6 tube
C. 100,000-ohm resistor

$$\phi_M(\text{flux in lines or maxwells})=\frac{\text{area of one-half of voltage wave in square inches}}{\text{number of turns on search coil}}\cdot\left(\frac{\text{volts}}{\text{inch}}\right)^{*}\cdot\left(\frac{\text{seconds}}{\text{inch}}\right)^{*}\cdot10^{8}$$

The maximum flux density B_m is found by dividing the flux by the area of steel that the coil encloses. This assumes that the flux is confined entirely to the steel or that the magnetic field in air is small compared to the magnetic field in the steel sheets.

$$B_m=\frac{\phi_m}{A}$$

where

A = area in square inches
B_m = maximum flux density in lines per square inch

A direct reading meter can be made to evaluate $\int_{t_1}^{t_2}e.dt$ or the average voltage. It was possible to measure this integral for determination of ϕ_m by using a bridge-type full-wave rectifier (Figure 22) and suitable d-c meter. The rectified direct voltage is approximately 1/20 of the alternating voltage input. A search coil of ten turns of number 22 wire was used with the instrument setup.

$$\phi_m=\frac{E_{avg}}{240N}\times10^8$$

where

$\frac{1}{240}$ = one-fourth of a cycle of a 60-cycle system
E_{avg} = average volts

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* These are calibrations measured from the oscillogram of the voltage wave.

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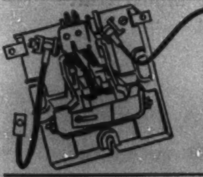

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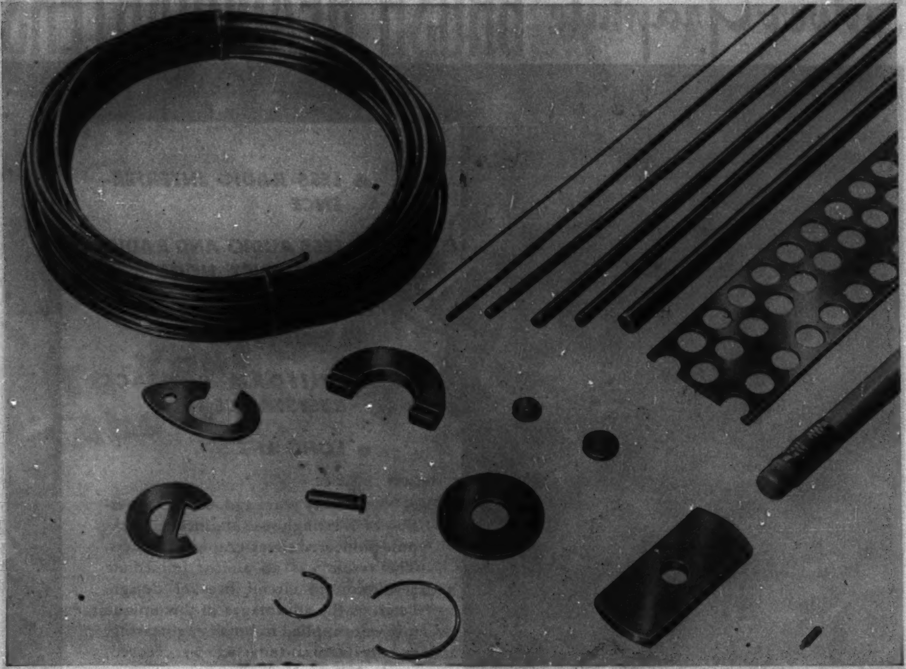

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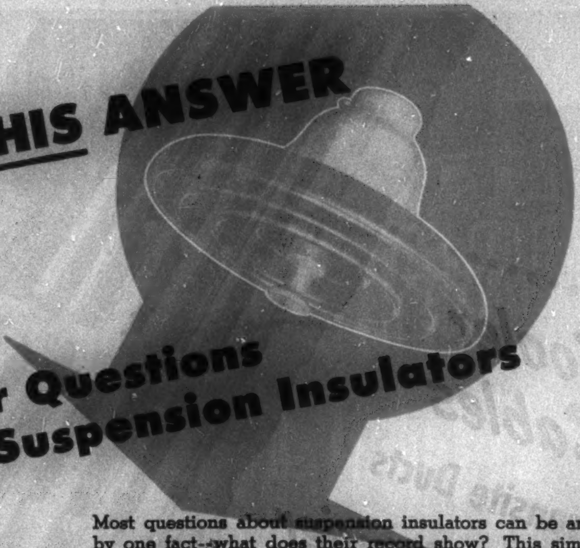

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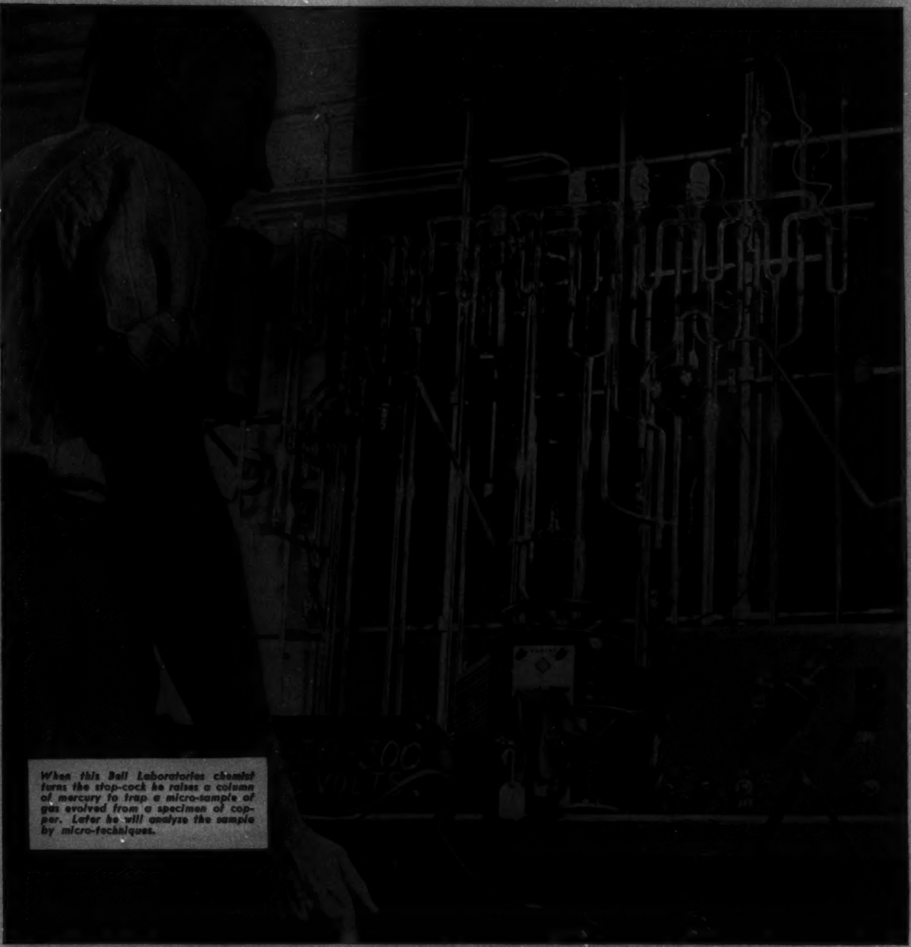

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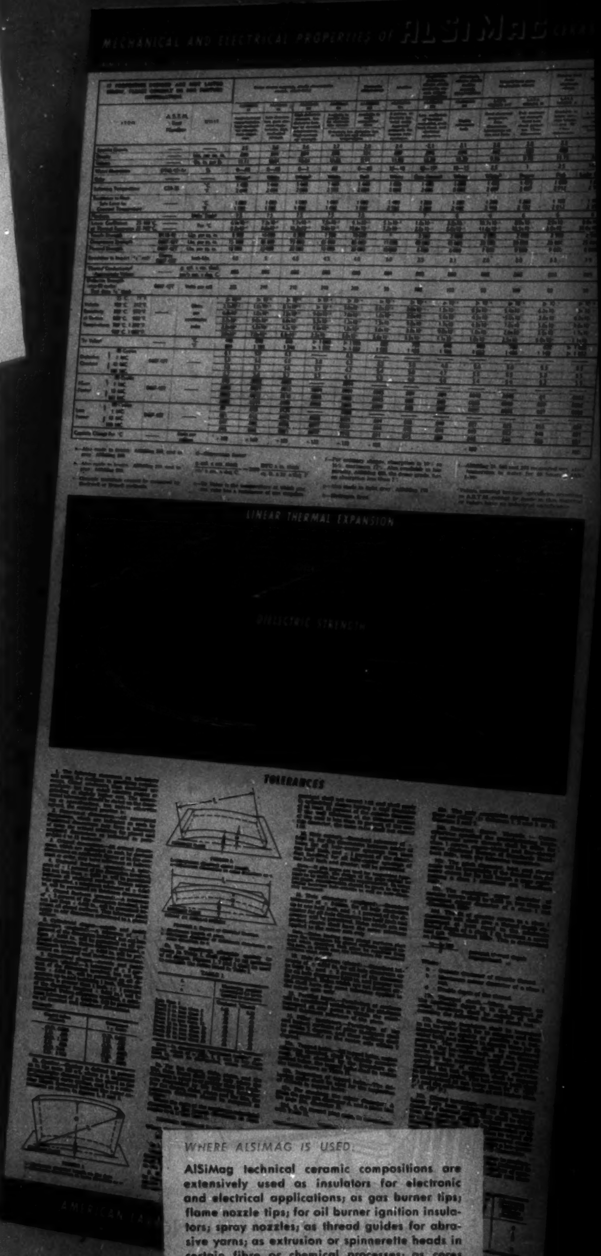
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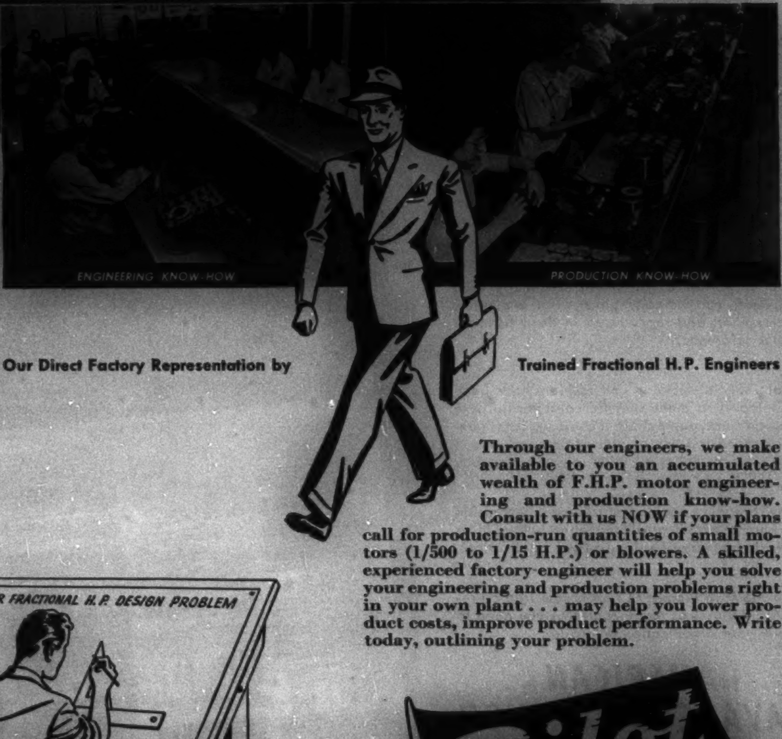

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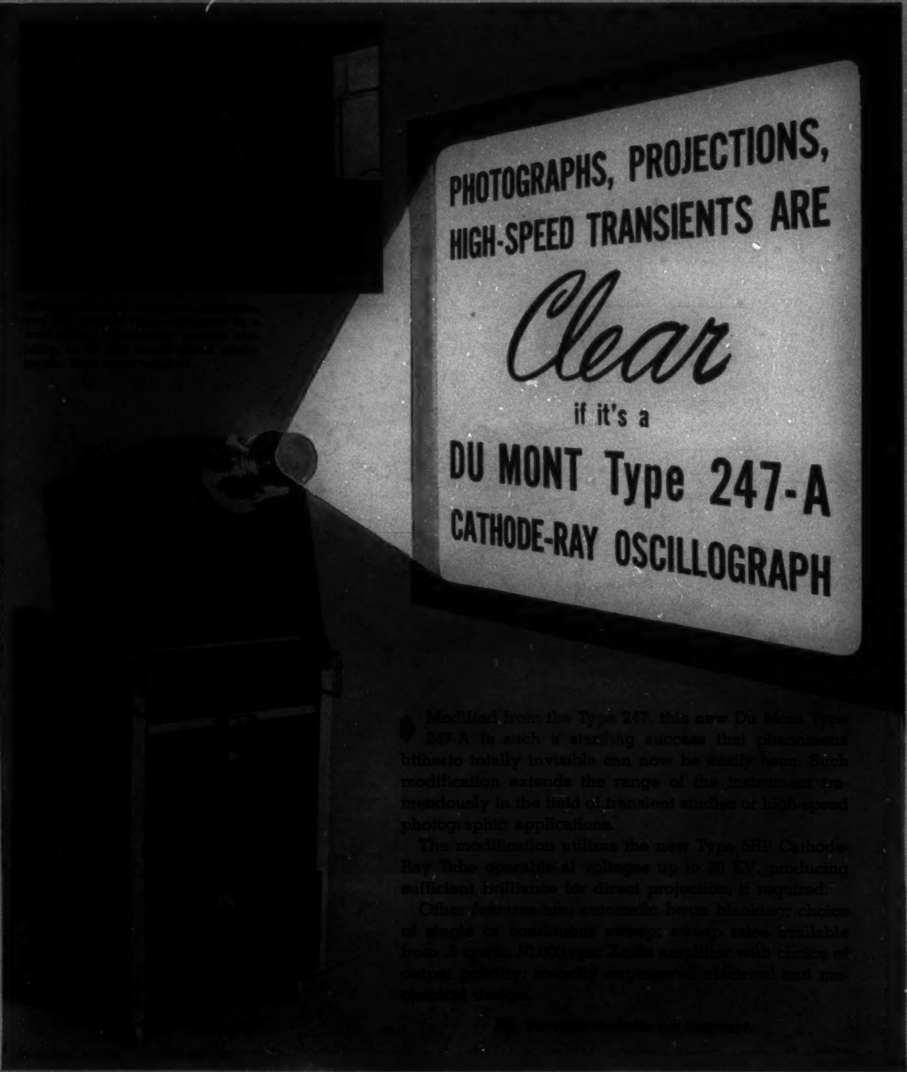
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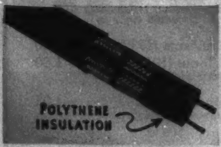

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Tensile Strength, lb. per sq. in. (dumbell specimen tested)	1600	1600
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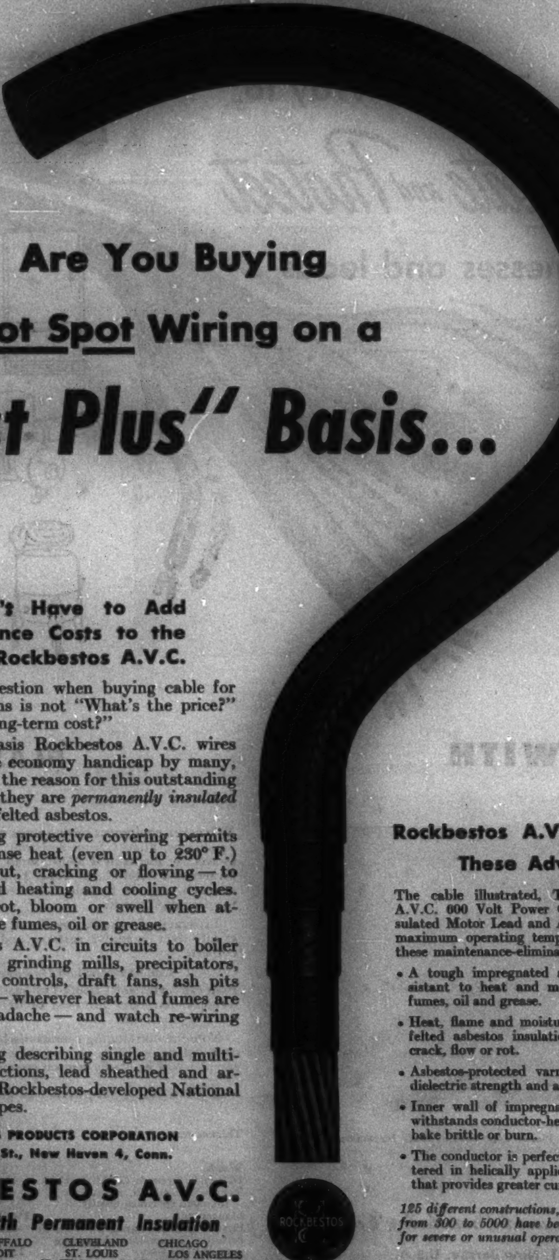

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Send for a catalog describing single and multi-conductor constructions, lead sheathed and armored, and many Rockbestos-developed National Electrical Code types.

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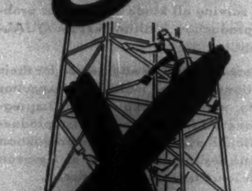

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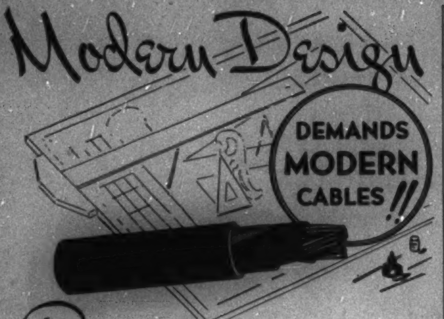

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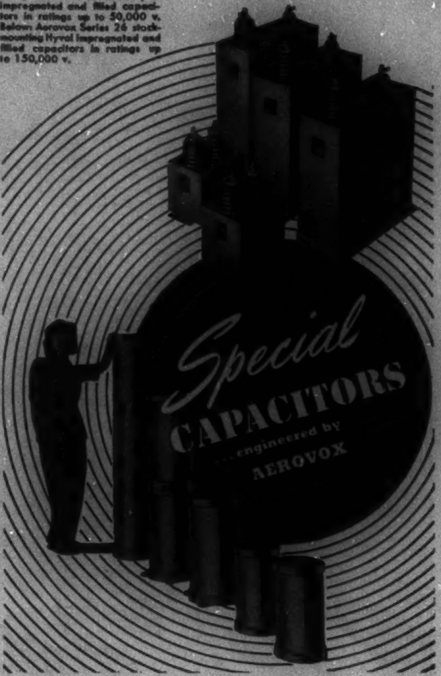

At right: Aerovox Series 20 Hyvol impregnated and filled capacitors in ratings up to 50,000 v. Below: Aerovox Series 26 stack-mounting Hyvol impregnated and filled capacitors in ratings up to 150,000 v.

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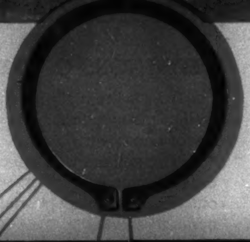

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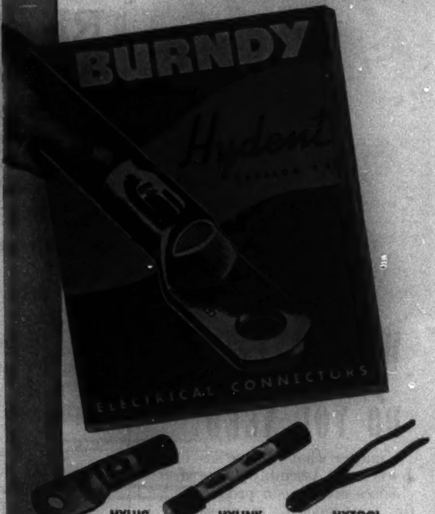

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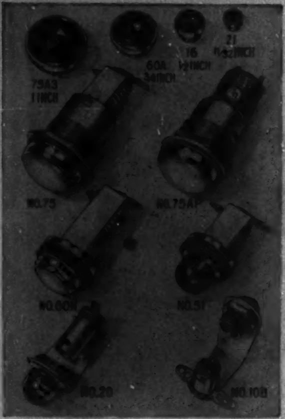

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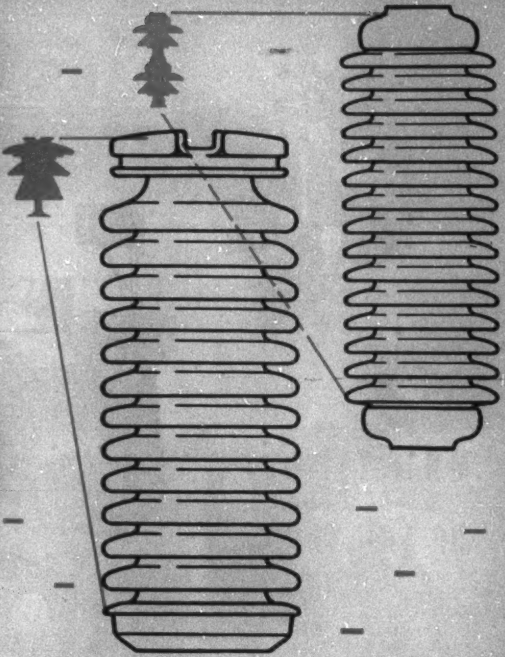

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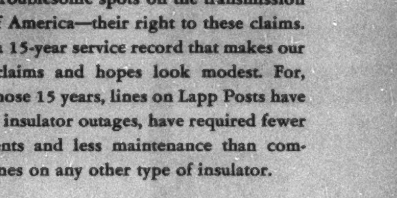

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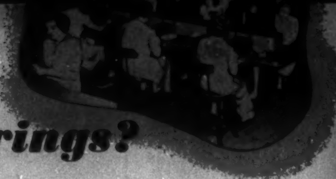

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We show you a few of the machines on these two pages, some mechanical, some electronically operated.

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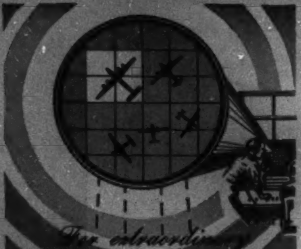

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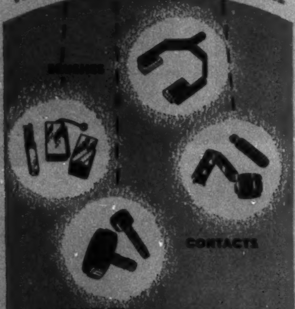

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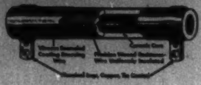

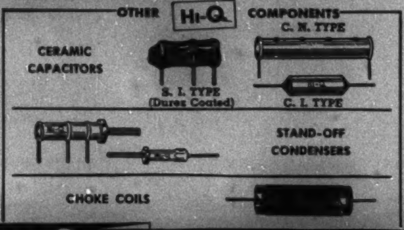

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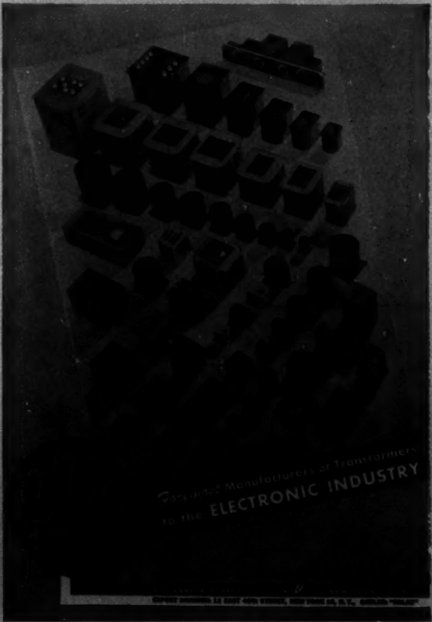

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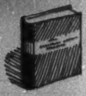

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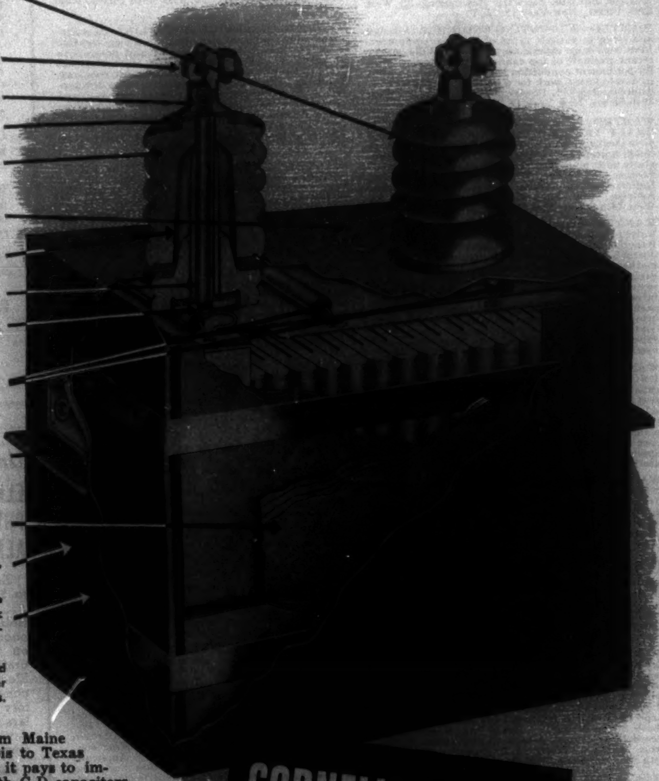

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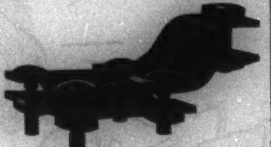

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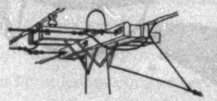

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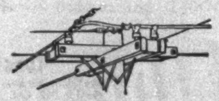

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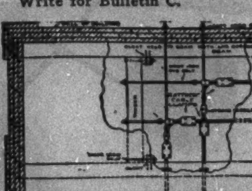

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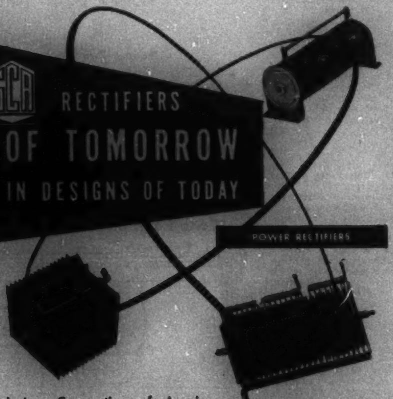

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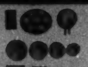

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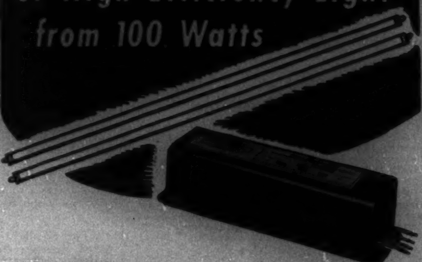

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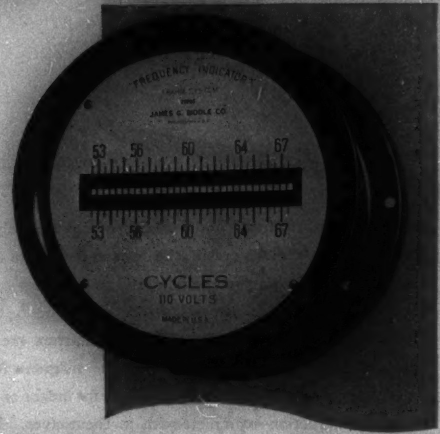

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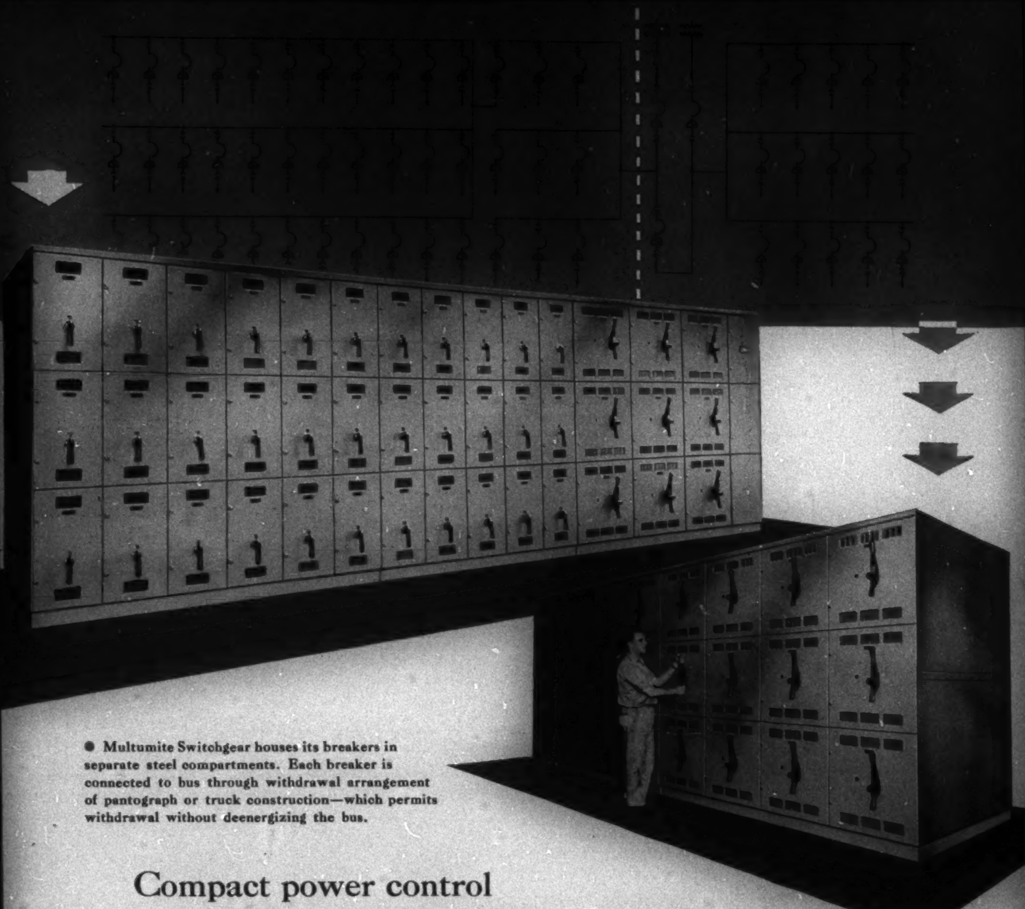

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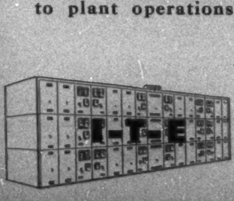

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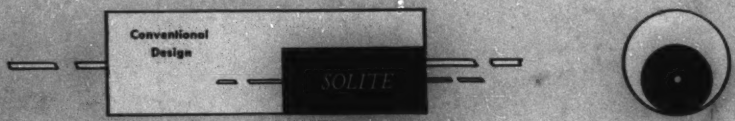

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0.5	200	1⅛	15/32	2	11/16	0.13	0.75
1.0	200	1⅛	23/32	2½	13/16	0.26	1.15
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SUPPLEMENT TO ELECTRICAL ENGINEERING TRANSACTIONS SECTION

PUBLISHED BY THE
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DECEMBER 1946

Supplement to Electrical Engineering—Transactions Section

December 1946

This supplement contains pages 861–1196, inclusive, of the 1946 AIEE Transactions, volume 65, published in this form for the convenience of members who do not subscribe to Transactions. It contains: (1) those technical papers presented at the 1946 AIEE North Eastern District meeting, summer convention, Pacific Coast convention, and Great Lakes District meeting, and not published previously in the monthly Transactions sections of Electrical Engineering; (2) all approved discussions on papers presented at the 1946 AIEE Southern District meeting, summer convention, and Pacific Coast convention.

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* Paper previously published.
† No discussion.

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Price—50 cents. No discount

Radio Telemetering for Testing Aircraft in Flight

CARL L. FREDERICK
MEMBER AIEE

Synopsis: A brief résumé is given of the factors which necessitated the development of thoroughly practicable and reliable telemetering equipment for dynamic testing of high-speed aircraft in flight. An outline of a radio telemetering system is presented which was developed under contract and accepted as suitable for structural flight testing by the United States Navy Bureau of Aeronautics in April 1945. Emphasis is placed upon the electric system and circuits which are of most interest to Institute members. These include the stabilized amplitude modulated subcarrier oscillators, the master amplifier, the air-borne regulated power supply, the frequency modulated transmitter, the automatic frequency controlled frequency modulated receiver, the heterodyne analyzer, and the mobile receiving station.

AN accurate knowledge of the amplitude, frequency, and relative phases of dynamic strains, pressures, and accelerations to be experienced by aircraft in flight is of vital importance in its design. In the past, much of this information could be calculated or determined by wind tunnel tests on small models. To meet the recent demands for speeds near the velocity of sound, or greater, it has become increasingly difficult, if not impossible, to extrapolate wind tunnel data. Means for telemetering this information from experimental aircraft and recording the data on the ground then became a necessity. With the advent of wind tunnels capable of testing powered models having spans up to 12 feet under conditions of pressure up to four atmospheres, and air velocities up to the speed of sound,[1] aerodynamic knowledge as to the effects of compressibility will be expanded greatly. However, the demand for even higher speeds makes the most modern wind tunnels incapable of furnishing all the required aerodynamic knowledge. A continued need for telemetering is therefore indicated, not only because it is the only practicable method of testing full scale remotely controlled aircraft designed to operate at nearsonic or supersonic speeds, but experience has indicated considerable savings in cost and manpower over other methods of testing larger and slower airplanes.

As often happens when research projects are undertaken to meet an urgent demand for one use, many other applications present themselves. Since the telemetering equipment has been completed, requests have come to use it in modified form for testing extremely high-speed compact flying bodies, for medical research to learn the effects on the human body when separated from high-speed aircraft flying at high altitudes, for testing and monitoring wind tunnels, and for observing and recording at the same time many items of information at remote points in large industrial plants or railroad yards.

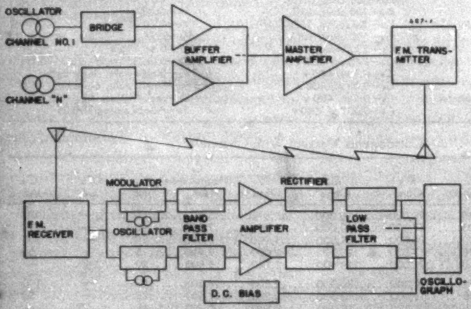


Figure 1. Block diagram of radio telemetering

As indicated, any number of channels (*N*) up to 14 may be included and simultaneously operated in the radio telemetering system

In planning the radio telemetering system herein described, established carrier telephone art was employed freely. As early as March 1916, R. A. Heising, of the Western Electric Company, filed an application for a patent on plural channel signaling. Numerous improvements have been made during succeeding years, and many of these are incorporated in the telemetering system designed by the Cornell Aeronautical Laboratory. The main point to be emphasized here is that this basic system was considered to be the most promising for development from the standpoint of proved reliability and simplicity, and the most likely to yield a practicable aircraft telemetering instrument with the minimum expenditure of time and manpower.

General Scheme of the Radio Telemetering System

This system may be described as an amplitude modulated subcarrier and frequency modulated radio carrier system as shown in Figure 1. The 14 subcarrier oscillators supply alternating current of frequencies from about 10 to 50 kilocycles to Wheatstone bridges incorporating strain-sensitive wires. These may be of the bonded-type or the unbonded-type gages with strain-sensitive wires incorporated in the instruments to measure pressures, accelerations, or forces. The variation in the resistance of the strain-sensitive wires changes the amplitude of the subcarrier oscillator. These changes from subcarrier channels are then amplified, collected, further amplified by the master amplifier, and then sent to the radio transmitter. The radio transmitter changes the amplitude modulated signals to a frequency modulated signal, the modulation of which is transmitted and detected by the radio receiver. Heterodyne analyzers are provided in a mobile receiving station to separate the individual signals and direct them to the

---

Paper 46-87, originally presented at the AIEE North Eastern District meeting, Buffalo, N. Y., April 24–26, 1946; recommended for publication by the AIEE committees on air transportation, communication, and instruments and measurements. Manuscript submitted March 6, 1946; made available for printing March 21, 1946.

Carl L. Frederick is vice-president of Fredric Flader, Inc., Air and Power Research Engineers, North Tonawanda, N. Y. He was formerly chief, physics department, Cornell Aeronautical Laboratory, Cornell Research Foundation, Inc., Buffalo, N. Y.

Contributions made by various members of the physics department of the Cornell Aeronautical Laboratory to the design of the components herein described are gratefully acknowledged. The author is particularly indebted to Karl D. Swartzel and Edward R. Dye, assistant department heads, for their supervision of the project and to C. F. Curtis, consulting engineer, Hazeltine Electronics Corporation, for his work on the automatic frequency control circuits of the frequency modulated receiver.

appropriate galvanometer in the recording oscillograph where 14 lines are drawn on sensitive paper ten inches wide.

Stabilized Oscillators, Bridge Circuits, and Buffer Amplifiers

The schematic diagram of the stabilized oscillators and bridge circuit is shown in Figure 2. As indicated previously, it is necessary that any required flight information be first converted to some form of electric change prior to its transmission to the ground receiving station. This conversion is effected by oscillator bridge units used in conjunction with either a bonded strain gauge attached to the structural member or an unbonded strain-sensitive element incorporated into the indicating mechanism of a flight instrument. The oscillator generates a signal whose amplitude, when introduced into a resistance bridge circuit containing the strain-testing element, is modulated by an amount proportional to the strain involved, that is, to actual flight data. The oscillator and bridge units are essentially the nerve centers of the entire radio telemetering system. Each consists of three separate components built into a single small aluminum chassis. These components are

1. An oscillator circuit, providing the subcarrier signal.
2. Two arms and necessary balance control of a resistance bridge circuit.
3. A buffer amplifier.

The most critical of the three components is the oscillator circuit itself. The nature of its use is such that it must maintain nearly constant amplitude and frequency for all operating conditions of voltage and temperature. Since the strain-gauge circuits will transmit information directly proportional to amplitude, the output voltage level of each oscillator must be held to close limits of stability, consistent with the accuracy of the data to be telemetered. In our case, a regulation of about one per cent was attained. This stability is achieved by a circuit in which the amplitude is controlled by a thermal element used as one arm of the oscillator feedback bridge. Thus, when an increase in output drives the bridge harder, unbalance is reduced tending to hold the oscillator output voltage constant. The over-all stability of the oscillator is accomplished through

1. The action of the hot filament balance lamps.
2. The control of ambient temperature within the definite limits of from 80 to 120 degrees Fahrenheit.
3. By the adjustment of the effective temperature coefficient of the resistor designated R_1.
4. By moderately close control of the plate and heater voltages.

The balance lamps adequately suppress the amplitude changes associated with supply voltage variations up to plus or minus ten per cent. Regulation of the 22-volt supply is adequate to control plate and heater supply within this range. The balance lamps, however, are not

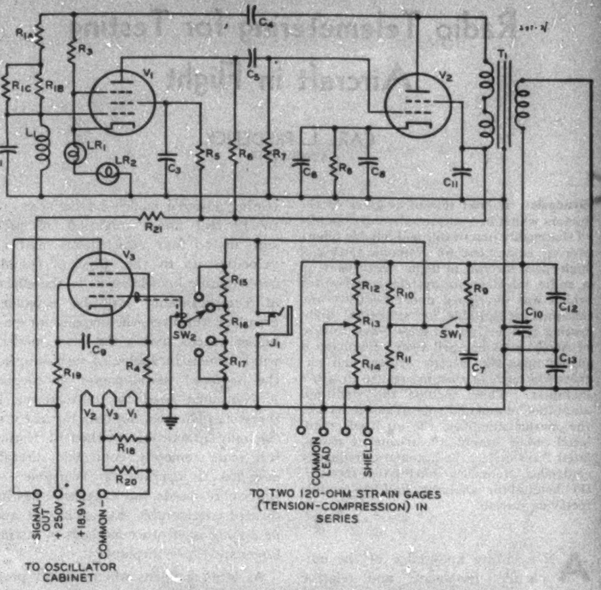


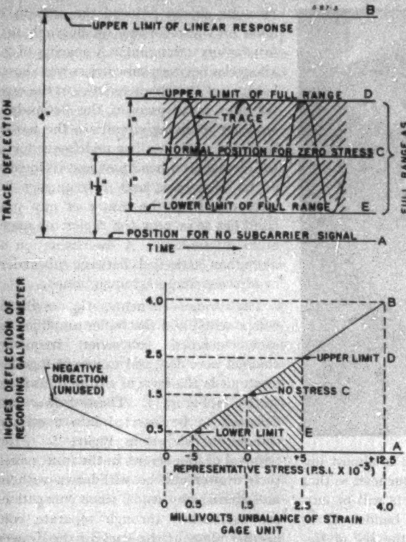
Figure 2. Schematic wiring diagram of oscillator and bridge balance unit

Components Common to All Channels

Symbol Description	Symbol Description	Symbol Description
R_{1C}—20,000 ohms, 1 w	R_{15}—1,250 ohms, 0.1%	C_9—0.05 μf, 400 v
R_4—1,250 ohms, 10%	R_{16}—750 ohms, 0.1%	C_{10}—140 μμf
R_5—1 megohm, 10%	R_{17}—500 ohms, 0.1%	C_{11}—0.25 μf, 400 v
R_6—0.1 megohm, 10%	R_{18}—100 ohms, 10%	V_1—6SJ7
R_7—0.5 megohm, 10%	R_{19}—1 megohm, 10%	V_2—6V6
R_8—750 ohms, 10%	R_{20}—500 ohms, 10%	V_3—6SJ7
R_9—48,000 ohms, 0.1%	R_{21}—250 ohms, 10%	LR_1—6 w, 115 v
R_{10}—120 ohms, 0.1%	C_3—0.25 μf, 400 v	LR_2—6 w, 115 v
R_{11}—120 ohms, 0.1%	C_4—0.1 μf, 400 v	SW_1—Single pole, single throw
R_{12}—2,000 ohms, 1%	C_5—0.05 μf, 400 v	SW_2—Single pole, 4-position
R_{13}—10,000 ohms potentiometer	C_6—0.5 μf, 400 v	J_1—Jack-Mallory A2A
R_{14}—2,000 ohms, 1%	C_8—0.001 μf, 600 v	T_1—Altec-Lansing TM205

Components Varying With Channels

Channel Number	Frequency Cycles Per Second	L_1 Megohms	R_{1A} Ohms	R_{1B} Ohms	R_3 Ohms	C_1 (Approximate) μμf	C_7 μμf	C_{12} μμf	C_{13} μμf
1	10,833	40	22,000	10,000	1,870	5,600	50	—	100
2	13,333	20	17,300	10,000	1,850	7,100	20	—	50
3	15,833	20	26,800	15,000	1,920	5,000	10	—	100
4	18,333	10	17,500	10,000	2,000	7,500	5	—	50
5	20,833	10	21,000	12,000	1,740	5,800	20	250	—
6	23,333	10	50,000	30,000	2,500	4,700	25	—	100
7	25,833	5	15,000	10,000	2,220	7,600	25	—	50
8	28,333	5	26,100	15,000	2,500	6,300	15	50	—
9	30,833	5	30,000	15,000	2,500	5,400	20	—	100
10	33,333	3	15,000	8,500	1,820	7,500	15	500	100
11	35,833	3	25,000	12,500	2,500	6,600	15	—	—
12	38,333	3	18,000	10,000	1,770	5,700	15	—	100
13	40,833	3	21,000	12,000	1,670	5,100	10	250	—
14	43,333	2	10,000	7,500	1,670	6,600	15	—	100



P-4-86
OSCILLATOR UNIT

Figure 4 (above). Oscillator unit

Figure 3 (left). Diagrams of signal ranges and the response of a recording galvanometer element to these changes

capable of stabilizing the changes in the tuned circuit resistance associated with the alteration in Q value of the coil L_1 caused by temperature changes. To overcome this difficulty, the 14 oscillator and bridge units are housed in a single cabinet whose temperature is held within the temperature limits mentioned. To prevent this last increment of amplitude change, a temperature compensating resistor consisting of a series-parallel combination of a highly negative coefficient resistor and two positive coefficient resistors are chosen so as to give an overall temperature coefficient which will supply the required compensation. The compensating network is mounted within the shield of the coil L_1 so that its effect will change at the same time rate as the Q of the coil. In theory it is possible to adjust for zero amplitude change for various temperatures. Control is not quite that accurate in practice; it has been possible to adjust and hold to less than two per cent change for a 40-degree Fahrenheit temperature rise permitted within the cabinet.

Because of the high Q of the tuned circuit, the frequency is relatively independent of plate and heater voltages. Through a plate voltage range of 250 plus or minus ten volts or for a change in heater voltage of ten per cent, the frequency change is less than two cycles per second. Frequency change with temperature is less than two cycles from 80 degrees to 120 degrees Fahrenheit because of the existing favorable temperature balance between the duo-lateral coils L_1 and the silver mica capacitors C_1.

Two arms of the strain-gauge resistance bridge are built into each circuit. The other two arms completing the bridge are formed by the strain-sensitive elements attached to the airplane. The necessary balance controls for both amplitude and phase adjustments must be included as part of the bridge unit. In this design it is imperative that drift of the bridge balance be prevented, since the indicated strain is proportional to the change in the bridge balance. The transformer T_1, connecting the oscillator to the strain-gauge circuit, has a balanced secondary to facilitate securing strain-gauge bridge balance and to reduce noise pickup. Two strain gauges attached to the structure under test are connected to the unit, one for receiving strain and the other as a temperature balancing element. The resistors R_{10} and R_{11} are fixed resistors built into the amplifier and oscillator assembly as discussed later. A potentiometer, R_{13}, is incorporated for balancing the resistive components to the bridge. R_{12} and R_{14} limit the total range of adjustment available on R_{13}. The capacitor C_{10} is for adjusting the reactive component of the bridge balance. As has been explained before, the carrier signal does not show readily the sense of the measured stress. For this reason the bridge is operated off-balance by an amount depending on the anticipated strain. At the receiving point, the nature of the response to the signal from a bridge adjusted in this manner is illustrated in Figure 3. The signal ranges are shown in the upper diagram with conservatively defined full range values and maximum linear response range. The lower diagram shows the response of a recording galvanometer element to the above signal

Figure 5. Telemetering cabinet, 14-channel
The rule is 18 inches long

P-5-4
TELEMETERING DRAWER
CABINET

Figure 6. Equipment installation in SB2A-4 airplane

ranges as received from a strain gauge connected to transmit stress information.

The principal considerations controlling the buffer amplifier design were that it must not cause cross modulation between various signals, and its gain must be held constant within extremely narrow limits. The buffer amplifier also serves the purpose of preventing undesirable coupling between the several strain-gauge circuits. Negative feedback reduces harmonic distortion and stabilizes the gain. The harmonic distortion of the circuit shown is less than 0.2 per cent with a 10-millivolt signal on the grid. The normal signal level is only 1.5 millivolts. During performance tests of this circuit, the gain varied two per cent as the plate voltage was changed from 200 to 300 volts and varied one per cent as the heater voltage varied from 5.5 to 7.0 volts. The 14 buffer amplifiers have a common plate feed resistor so that their signals are all added together in the common output circuit.

In selecting the oscillator frequencies, it was found in preliminary tests that distortion was relatively large in the frequency region below ten kilocycles. Therefore the subcarrier oscillator frequencies were all set above ten kilocycles, as shown in the table in Figure 2. The frequency of 10,833 cycles per second was chosen as a starting point so that an inharmonic relationship would exist between subcarriers, but noise caused by other modulation products may be just as serious as that caused by harmonics of subcarrier frequencies. It is almost impossible to allocate 14 frequencies so that all the modulation products will be outside of the transmission bands. However, it was found that the set of frequencies chosen reduced the amount of noise on the recording oscillograph to a satisfactory minimum. A spacing of 2.5 kilocycles between subcarriers was chosen because of the characteristics of the crystal filter units used in the heterodyne analyzer receiving circuits in the mobile receiving station. The mid-frequency of the crystal filter band pass units is located at 92 kilocycles, and a frequency 2.5 kilocycles from the center of any pass band lies at or near the region of maximum attenuation. A discrimination of more than 50 decibels between subcarriers of adjacent channels is maintained.

The bridge elements, the oscillator components, and the buffer amplifier for each assigned subcarrier frequency channel were designed to mount together as a unit in the form of a drawer, as illustrated in Figure 4. These drawers are designed to be inserted into a cellular cabinet, as shown in Figure 5. After removal of four screws in the front panel, each drawer can be withdrawn without unsoldering any leads, since connections are established through separate connecting plugs at the back of the drawer

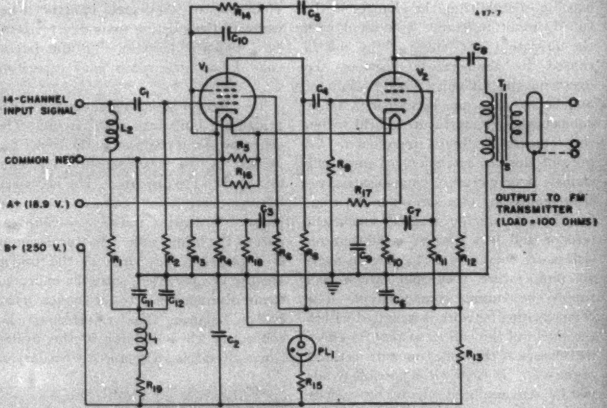


Figure 7. Schematic wiring diagram of master amplifier

Components Common to All Channels

Component	Description	Component	Description	Component	Description
V_1	6SJ7	R_{11}	0.15 megohm, IRC BT-1	C_5	0.01 μf
V_2	6V6, GTIG	R_{12}	0.015 megohm, IRC BT-1	C_6	0.5 μf
P_1	AN 3102-14S-2P (reworked)	R_{13}	2,250 ohms, IRC BW-1	C_7	0.1 μf
S_1	AN3102-14S-75	R_{14}	0.075 megohm, IRC WW-4	C_8	0.01 μf
R_1	0.01 megohm, IRC* ABA	R_{15}	0.5 megohm, IRC BT-1	C_9	0.1 μf
R_2	0.5 megohm, IRC BT-1	R_{16}	250 ohms, IRC BW-1	C_{10}	5 μf
R_3	2,250 ohms, IRC BW-1	R_{17}	15 ohms, IRC ABA	C_{11}	0.5 μf
R_4	2,250 ohms, IRC BW-1	R_{18}	0.01 megohm, IRC ERIE	C_{12}	0.5 μf
R_5	50 ohms, IRC ABA	R_{19}	0.001 megohm IRC BW-1	L_1	40 mh, Delta R.**
R_6	0.5 megohm, IRC BT-1	C_1	0.01 μf	L_2	30 mh, Delta R.
R_7	0.05 megohm, IRC BT-1	C_2	0.5 μf	PL_1	1/25-w neon lamp
R_8	0.5 megohm, IRC BT-1	C_3	0.25 μf	T_1	Altec*** TM205
R_{10}	200 ohms, IRC BW-1	C_4	0.0001 μf		

*International Resistance Company. ** Delta Radio Corporation. *** Altec Lansing Corporation.

Each drawer weighs 2.75 pounds. The cabinet with drawers mounted, including the master amplifier and the temperature controlling unit, weighs 60 pounds and occupies 2.25 cubic feet. The cabinet, together with its associated voltage regulator, power supply, and frequency modulated transmitter, is shown in Figure 6 installed in the rear observer's cockpit of the SB2A-4 airplane.

Master Amplifier

The function of the master amplifier, shown schematically in Figure 7, is to receive and superpose the comparatively weak signals from the 14 buffer amplifiers, described previously, and to raise the composite signal level to an optimum value for input to the frequency modulated transmitter. The 2-stage resistance-capacitance coupled amplifier with a large negative feedback provides 17 ±0.5 decibel of gain with less than 0.25 per cent total distortion over the frequency range from 10 to 50 kilocycles.

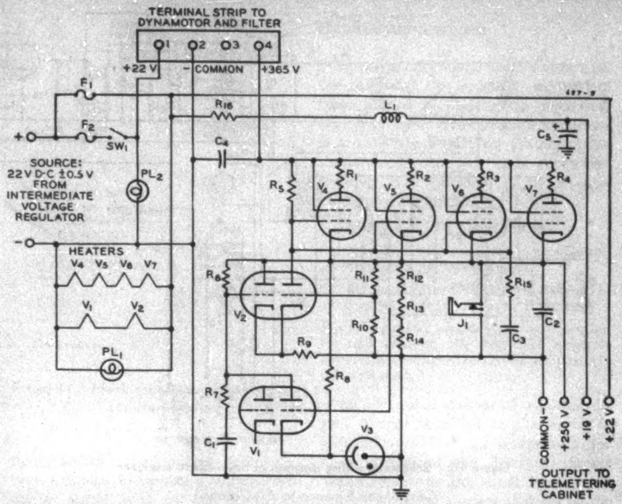


Figure 8. Schematic wiring diagram of air-borne power supply

Components Common to All Channels

Component Description
R_1, R_2,
R_3, R_4—100 ohms
R_5—0.3 megohm
R_6—1.5 megohms
R_7—5,000 ohms
R_8—10,000 ohms
R_9, R_{10}—0.1 megohm
R_{11}—0.15 megohm
R_{12}—0.25 megohm
R_{13}—0.05 megohm
R_{14}—0.1 megohm
R_{15}—25,000 ohms
R—Ten 5-ohm, 10-w resistors parallel
C_1—0.013 μf

Component Description
(0.01 and 0.003 in parallel)
C_2, C_4—2.00 μf, 600 v
C_3—0.00075 μf
C_5—25 μf, 50 v
V_1, V_2—12SL7GT
V_3—VR-75
V_4, V_5, V_6, V_7—6Y6G
PL_1, PL_2—28-v pilot lamps
SW_1—Single pole, single throw, 30 amps, 28 v
J_1—Standard phone jack
F_1—20 amps, Littelfuse 4AG
F_2—40 amps, Littelfuse 4AG
L_1—100 mh

Air-Borne Regulated Power Supply

The accuracy and stability of the air-borne radio telemetering equipment depend upon a high degree of stability of the plate supply. It was therefore necessary to develop an extremely well-regulated 250-volt source to supply the vacuum tubes in the oscillator cabinet. As is well known, the voltage supply by an airplane under test may vary from 20 to 30 volts depending upon the maneuver. An intermediate-voltage regulator was designed using a series of extremely fast-acting relays to supply a primary power source of 22 plus or minus 0.5 volt direct current for the air-borne regulated power supply. Space will permit only a brief description of the air-borne regulated power supply, shown schematically in Figure 8, and its regulation characteristics in Figure 9. The circuit is a direct-coupled amplifier having a gain of 55 decibels and a negative feedback of 44 decibels. The input to the amplifier is derived by balancing a portion of the output voltage against a

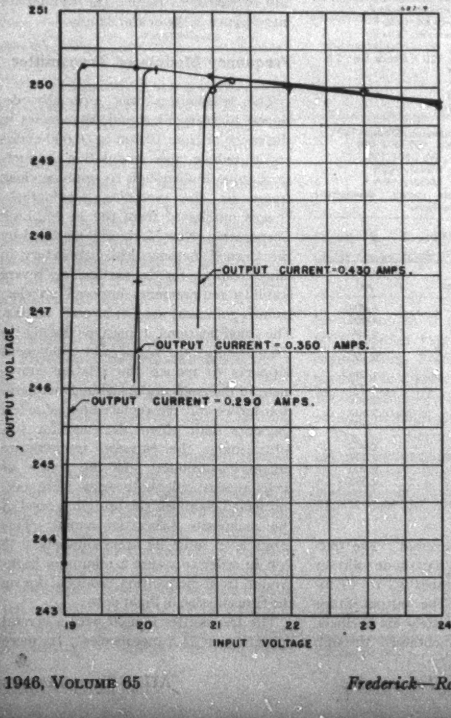


Figure 9. Regulation characteristics of air-borne power supply

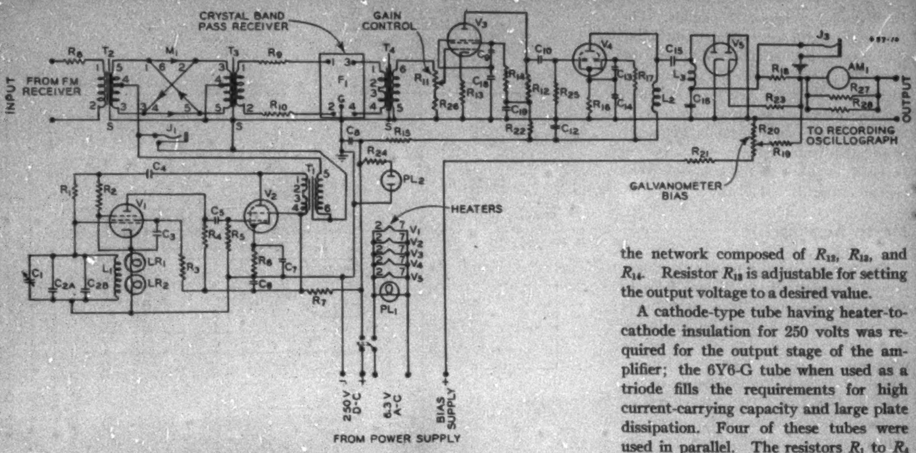


Figure 10. Schematic wiring diagram of heterodyne analyzer

Components Common to All Channels

Component	Description	Component	Description	Component	Description
R_1*	IRC BT-1	R_{23}	1,000 ohms, IRC BT-1	C_{11}	0.1 μf, 400 v, paper
R_2	2,500 ohms, IRC BW-1	R_{24}	0.5 megohm, IRC BT-1	C_{12}	0.05 μf, 400 v, paper
R_3	0.5 megohm, IRC BT-1	R_{25}	0.5 megohm, IRC BT-1	C_{13}	0.05 μf, 400 v, paper
R_4	0.025 megohm, IRC BT-1	R_{26}*	IRC BT-1	C_{14}	0.05 μf, 400 v, paper
R_5	0.5 megohm, IRC BT-1	R_{27}	50 ohms, IRC BW-1	C_{15}	0.05 μf, 400 v, paper
R_6	750 ohms, IRC BW-1	R_{28}	150 ohms, IRC BW-1	C_{16}	0.05 μf, 400 v, paper
R_7	1,000 ohms, IRC BW-2	T_1	Specifications, Altec TM206	C_{17}	0.1 μf, 400 v, paper
R_8	600 ohms, IRC WW-3	T_2	WE 146F	L_1	0.5 megohm, Delta R.
R_9*	150 ohms, IRC WW-3	T_3	WE 146F	L_2	20 megohm, Delta R.
R_{10}*	150 ohms, IRC WW-3	T_4	WE 151B	L_3	40 megohm, Delta R.
R_{11}	0.03-megohm potentiometer, wire wound	PL_1	6.3-v 0.15-amp pilot lamp	LR_1	6 w, 120 v, Mazda lamp S-6
R_{12}	0.2 megohm, IRC BW-1	PL_2	1/25-w neon lamp	LR_2	6 w, 120 v, Mazda lamp S-6
R_{13}	1,000 ohms, IRC WW-4	C_1†	Variable, Budd	V_1	6SJ7
R_{14}	1 megohm, IRC BT-1	C_2A	Silver mica	V_2	6V6
R_{15}	1,000 ohms, IRC BW-2	C_2B‡	Silver mica	V_3	6SJ7
R_{16}	500 ohms, IRC WW-4	C_3	0.05 μf, 400 v, paper	V_4	6V6
R_{17}	0.02 megohm, IRC BT-1	C_4	0.01 μf, 300 v, mica	V_5	6H6
R_{18}	1,000 ohms, IRC WW-4	C_5	0.05 μf, 400 v, paper	AM_1	Milliammeter, Simpson 127
R_{19}	1,000 ohms, IRC WW-4	C_6	0.1 μf, 400 v, paper	M_1	Varistor, WE D97 428
R_{20}	2,000-ohm potentiometer, wire wound	C_7	25 μf, 50 v, electrolytic	F_1	Filter, WE D168 159
R_{21}	0.02 megohm, IRC WW	C_8	0.1 μf, 400 v, paper	J_1	Jack, Mallory A2A
R_{22}	0.01 megohm, IRC BT-1	C_9	0.05 μf, 400 v, paper	J_2	Jack, Mallory A2A
		C_{10}	0.0005 μf, mica	SW_1	Toggle switch, double-pole single-throw

* Adjust according to test specifications. † See the following table. ‡ As determined by text.

Channel Number	Frequency, Cycles Per Second	C_1 Max, μμf	C_2 (Approximate), Total, μμf	R_1 (Approximate)	C_2A, μμf
1	102,833	100	4,800	25,000	4,000
2	105,333	100	4,500	26,400	4,000
3	107,833	100	4,300	27,500	3,500
4	110,333	100	4,200	28,800	3,500
5	112,833	100	4,000	29,800	3,500
6	115,333	100	3,900	30,700	3,500
7	117,833	100	3,700	32,000	3,000
8	120,333	50	3,500	33,000	3,000
9	122,833	50	3,400	34,400	3,000
10	125,333	50	3,300	35,400	3,000
11	127,833	50	3,100	36,400	2,500
12	130,333	50	2,900	37,900	2,500
13	132,833	50	2,800	39,000	2,500
14	135,333	50	2,700	40,300	2,000

constant d-c potential of 75 volts determined by the gas-filled tube V_3. The first stage of amplification makes use of one triode section of V_1 and has a gain of 30 decibels. The second stage uses both sections of V_2 to obtain a gain of 30 decibels with no phase reversal. The output of the amplifier is a cathode follower stage using vacuum tubes V_4 to V_7 inclusive in parallel. The output stage has a loss of approximately six decibels. Negative feedback is obtained through the network composed of R_{12}, R_{13}, and R_{14}. Resistor R_{13} is adjustable for setting the output voltage to a desired value.

A cathode-type tube having heater-to-cathode insulation for 250 volts was required for the output stage of the amplifier; the 6Y6-G tube when used as a triode fills the requirements for high current-carrying capacity and large plate dissipation. Four of these tubes were used in parallel. The resistors R_1 to R_4 inclusive, in series with the plate leads, suppress any high frequency parasitic oscillations.

Power for the vacuum tube heater circuits in the oscillator cabinet (seven amperes direct current at 19 volts) is obtained from the 22-volt input through a series dropping resistor R_{18}. The coil L_1 and capacitor C_3 reduce the dynamotor interference in the heater supply.

Frequency Modulated Transmitter

The transmitter was especially designed to transmit signal frequencies in the range of from 10,000 to 50,000 cycles and to reduce cross modulation products to a value low enough to produce clean traces on the recording oscillograph. It was purchased from the F. M. Link Corporation, New York, and modified by the Cornell Aeronautical Laboratory in minor respects for the particularly severe stability requirements imposed on electronic equipment used in aircraft testing. The modifications consisted mainly in strengthening the frame and component supports to reduce the relative movement of the parts under the high acceleration forces and the replacement of a few elements with others designed for stability under the extreme temperature ranges encountered. In the course of development, adequate tests were performed to establish the optimum level of the composite subcarrier signal. This input level must be approximately 0.35 volt in order to assure a minimum background from modulation products during the transmission of flight data.

The transmitter is used with a crystal operating on 71.5 megacycles. Its more

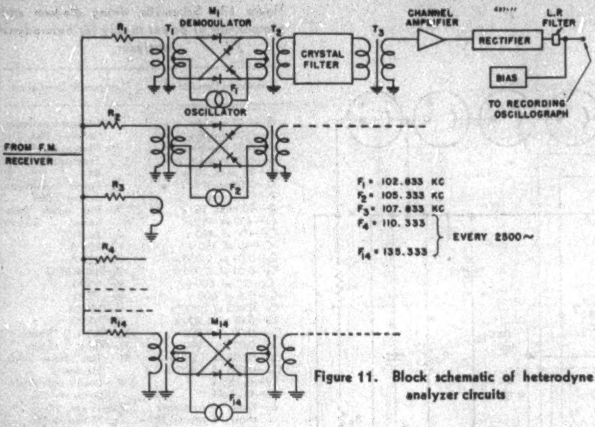


Figure 11. Block schematic of heterodyne analyzer circuits

important characteristics are as follows:

Power output	50 watts
Frequency range	65 to 72 megacycles
Frequency deviation	±150 kilocycles
Audio response	±1 decibel 1,000–50,000 cycles per second (overall; transmitter-receiver)
Modulation stability	Within one per cent for primary source variation between 21.5–22.5 volts
Mounting	Shock mounted on base
Size	$16^1/_8$ by $8^1/_4$ by $7^7/_8$ inches
Power required	7.5 amperes at 28 volts direct current
Output impedance	50–100 ohms into concentric line

The transmitter is self-contained with its dynamotor power supply. Both filament and dynamotor currents are obtained through the remote control box which is equipped with separate switches and pilot lights for each. A metering jack on the front panel permits measuring the audio input level. A meter and selector switch also are provided on the front panel to permit measuring grid currents in all necessary stages when making adjustments.

Automatic Frequency Controlled Frequency Modulated Receiver

The frequency modulated receiver used was purchased from the F. M. Link Corporation, New York, N. Y., and modified for automatic frequency control. The automatic frequency control feature is not required for many applications where precise crystal control may be obtained in the transmitter. However, the requirements on the frequency stability of the transmitter may be greatly reduced if automatic frequency control is used in the receiver. (In fact, crystal control of the transmitter is not feasible in many applications of telemetering for compact flying bodies.) The output circuit has been modified to furnish a suitable subcarrier signal level to the heterodyne analyzer circuits.

The receiver may be adjusted manually over the approximate range of 66 to 74 megacycles and automatically over the range of plus or minus 0.5 megacycle from the manual setting. It has been designed especially to receive signal frequencies in the range of 10,000 to 50,000 cycles with sufficient fidelity of reproduction, holding cross modulation products to a value low enough to permit the recording of clean traces on the recording oscillograph paper.

The more important characteristics of the modified receiver are

Frequency range	65 to 72 megacycles
Type of signal	Frequency modulated ±150 kilocycles maximum deviation
Audio response	±1 decibel, 10,000–50,000 cycles per second (overall; transmitter-receiver)
Output stability	Within one per cent for primary source variation between 24–28 volts
Power required	3 amperes at 28 volts direct current
Output impedance	100 ohms
Output power	1/2 watt maximum

Heterodyne Analyzer

The composite signal which appears at the output of the frequency modulated radio receiver is analyzed, or separated into its 14 separate components, by means of a heterodyne analyzer system of 14 similar units, one for each subcarrier channel. The principal parts of each unit shown in Figure 10 are

1. A copper-oxide ring modulator.
2. A vacuum tube oscillator to provide the best frequency for the subcarrier.
3. A crystal filter for selecting the modulated signal on that subcarrier frequency.
4. An amplifier and detector combination for each channel, the output of which is applied to its assigned recording-oscillograph galvanometer.

An adjustable amount of regulated direct current is added to the signal as galvanometer bias to compensate for that portion of the detected current which represents the signal level of the subcarrier in the absence of strain on the gage at the transmitter. A milliammeter in series with the output gives visual indication of the signal level.

The block schematic shown in Figure 11 will serve to present the general plan of the analyzer. The 14 modulated subcarriers appear on the primary winding of the transformers, T_1. It was considered advisable to use identical crystal filters for each channel because procurement was simplified and also because changes in subcarrier frequencies were facilitated when desirable. Frequencies for the 14 local oscillators then were chosen to permit the desired side band to pass through the filters as shown. For example, the local oscillator frequency of F_1 is chosen of the value 102.833 kilocycles to apply to demodulator M_1. The subcarrier T_1 of frequency 10.833 kilocycles then combines with F_1 to produce a difference frequency of 92.000 kilocycles which passes through the crystal filter. It will be noticed that the attenuation characteristic shown in the insert of

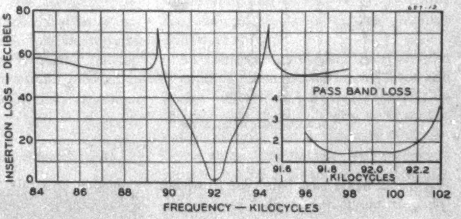


Figure 12. Crystal filter characteristic

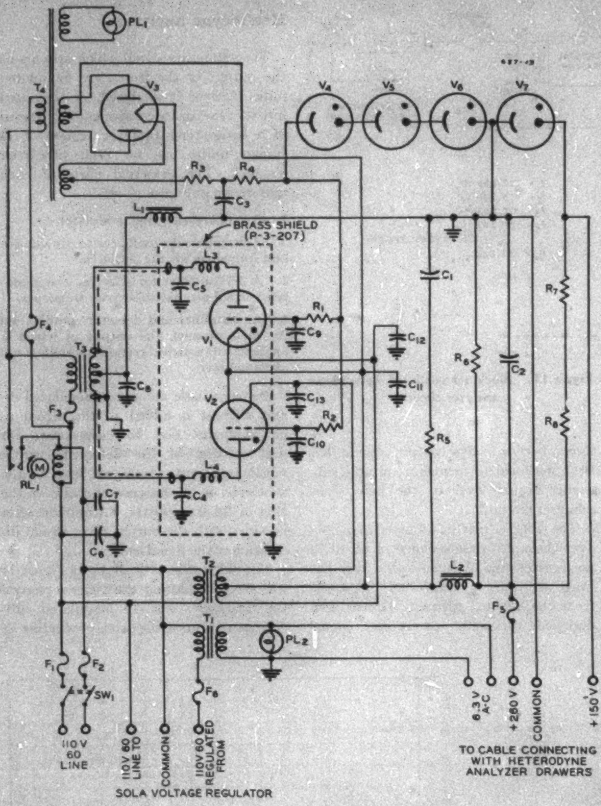


Figure 13. Schematic wiring diagram and components of power supply for heterodyne analyzer

Component Description

R_1—0.05 megohm, 1 w
R_2—0.05 megohm, 1 w
R_3—0.01 megohm, 10 w
R_4—0.01 megohm, 2 w
R_5—10 ohms, 10 w
R_6—0.02megohm, 25 w
R_7—750 ohms, 25 w
R_8—50 ohms, 10 w
C_1—400 μf, 450 v
C_2—400 μf, 450 v
C_3—40 μf, 450 v
C_4—0.01 μf, 2,000 v
C_5—0.01 μf, 2,000 v
C_6—0.1 μf, 600 v
C_7—0.1 μf, 600 v
C_8—0.005 μf,1,200 v
C_9—0.02 μf, 1,200 v
C_{10}—0.02 μf, 1,200 v
C_{11}—0.01 μf, 1,200 v
C_{12}—0.005 μf, 1,200 v
C_{13}—0.005 μf, 1,200 v
L_1—4h choke
L_2—4h choke
L_3—Radio frequency choke, 1 amp (P-1-055)
L_4—Radio frequency choke, 1 amp (P-1-505)

Component Description

T_1—Filament transformer, Jeffries 1,290
T_2—Filament transformer, Thordarson T-19F-84
T_3—Power transformer, Jeffries 1,288
T_4—Bias supply transformer, Stancor P6,010
V_1—393 A
V_2—393 A
V_3—5W4GT-G
V_4—VR90
V_5—VR90
V_6—VR105
V_7—VR150
PL_1—6.3 v, 0.15 amp
PL_2—6.3 v, 0.15 amp
RL—Time delay relay, Hayden
SW—Double-pole single-throw switch
F—15 amp, 4AG
F_2—15 amp, 4 AG
F_3—5 amp, 4 AG
F_4—1 amp, 4AG
F_5—2 amp, 4 AG
F_6—5 amp, 4AG

Adjust time delay of RL_1 to 50 seconds ± 10 seconds. Connect capacitors C_9, C_{10}, C_{11}, C_{12}, and C_{13} as close as possible to sockets of V_1 and V_2.

Figure 12 has a pass band flat to within plus or minus 0.1 decibel over a frequency range of at least 200 cycles per second. The operator of the receiving station readily can adjust the oscillators to attain uniform attenuation and phase transmission of all frequencies represented in the modulation of the air-borne subcarrier oscillators produced by the instruments from which data are desired.

Each modulator unit, shown in Figures 10 and 11, consists of four copper-oxide rectifier elements arranged in a lattice structure. The input signal is applied through a transformer from the 1-ohm impedance in the output of the frequency modulated receiver. It is therefore attenuated sufficiently in passing through the input resistor R_1 of 600 ohms and the primary impedance of the transformer T_1, which is also 600 ohms, to prevent appreciable crosstalk from the modulator in a specified channel to the other analyzers which are connected in parallel. Since the input of each signal channel to the modulators is about 50 millivolts rms, the composite signal amplitude of the 14 channels is about 700 millivolts, which is about ten decibels below the 1-volt output of the oscillators. Increase in the signal level may result in overmodulation and increase the noise level of the oscillator output. The temperature characteristics of the copper-oxide rectifiers in the modulators are such that temperature regulation beyond that provided for the comfort of the operators of the mobile receiving station is not required.

Since the voltage level of the crystal filter output is low, 2-stage resistance-coupled amplifiers are employed in each channel as shown. There is sufficient variation in the performance of commercial 6SJ7 tubes to cause considerable difference in the crystal filter response curve for different volume control settings in the first amplifier grid circuit. By selecting for use only tubes with closely similar characteristics, this difficulty has been minimized.

To obtain the envelope containing the signal from the 92 kilocycles amplitude

Figure 14. Exterior view of mobile receiving station for radio and television telemetering

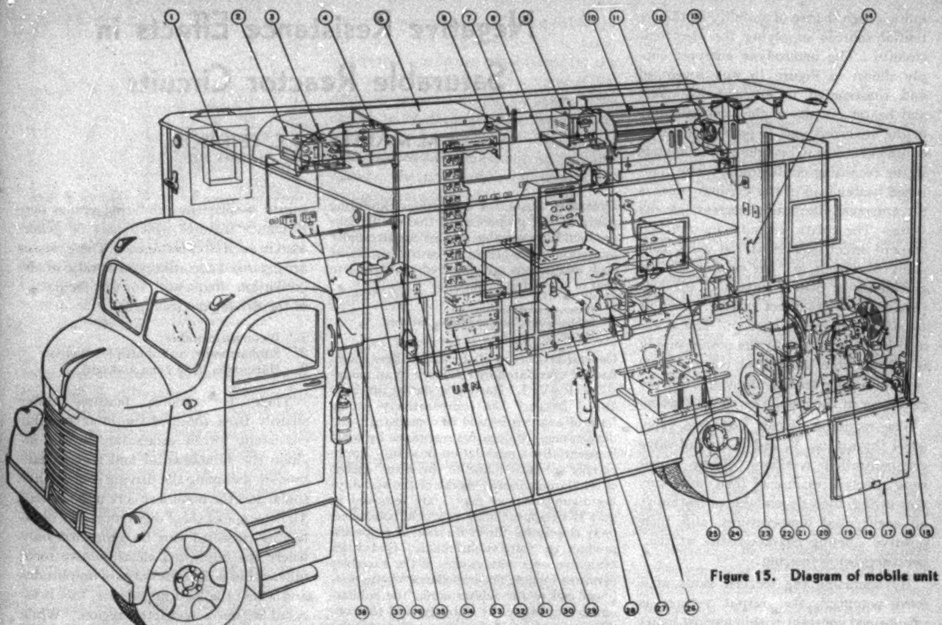


Figure 15. Diagram of mobile unit

modulated wave received from the amplifier, a 6H6 diode detector is used in a bridge-type half-wave detector circuit balanced to reduce the effect of variations in filament voltage. The carrier frequencies present in the rectifier output are suppressed by a low-pass filter system with an adequate attenuation above 300 cycles per second. Filters have been designed to reduce the frequency response of the telemetering system to any desired value. For example, in studying certain types of low frequency flutter, higher frequencies sometimes referred to as "aerodynamic hash" may be eliminated from the record, thus facilitating the analysis of the particular quantity under study.

Figure 16. Interior of mobile unit

As mentioned previously, an a-c bias is applied in the subcarrier oscillator and bridge unit for the purpose of identifying the direction or sense of change in strain-gauge circuits connected with the airborne radio transmitter. Zero strain is represented by a signal of half strength, plus strain by more than half, and minus strain by less than half. To remove the effect of this bias at the receiving station, a d-c bias obtained from the heterodyne analyzer power supply is applied to the galvanometers which record the signals received from the rectifiers. Since the 150 volts direct current from the power supply is constant within ±0.2 per cent, the galvanometers can be adjusted accurately and permanently for zero on the scale. The bias current is attenuated so that a current of zero to five milliamperes is available to balance the detected zero level signal. Galvanometer sensitivity is such that a 1-inch deflection results from about 67 per cent modulation.

Heterodyne Analyzer Power Supply

Accuracy and dependability of operation of the heterodyne analyzer unit require a high degree of stability in the potential sources supplying the electronic circuits. The heterodyne analyzer supply shown in Figure 13 was developed and constructed to energize the plate and heater circuits of the vacuum tubes in the analyzers and to provide a source of direct current to bias the galvanometers in the recording oscillograph. The combined heater load of the vacuum tubes is 27 amperes alternating current at 6.3 volts. The plate circuits require 1.0 ampere direct current at 250 volts, constant to within ±1 per cent, while for galvanometer bias a current of 0.1 ampere at 150 volts direct current, constant to ±0.2 per cent, is needed. The available input potential to the power supply is 115 ±5 volts alternating current at 60 ±5 cycles per second, derived from the generator in the mobile receiving station.

The regulator circuit uses two type-393A thyratrons, V_1 and V_2, as full-wave grid-controlled rectifiers. These tubes were selected because of their large current rating, their dependable operation at all anticipated temperatures, and their positive starting characteristics at temperatures below freezing.

By operating the rectifier grids at a fixed potential, the output voltage is maintained constant within narrow limits regardless of variations in input potential or load current. The fixed grid potential is obtained from three voltage-regulator tubes connected in series, V_4, V_5, and V_6. An auxiliary power supply composed of T_4, V_3, R_3, and C_3 provides sufficient potential to ionize the *VR* tubes and maintain fairly constant current through them.

A filter is used to smooth the ripple content in the rectifier output. It is important to note that the first filter choke, L_1, is in the negative lead, thus reducing the ripple in the voltage appearing between the rectifier cathodes and the point which fixes their grid potential.

Mobile Receiving Station

The mobile receiving station, shown in Figures 14, 15, and 16, was designed for efficient and reliable use in the field. Large savings in cost of flight testing can be effected if time can be saved during field tests because a large crew is generally involved, and flying time is very costly. Provisions were made for a specially designed 5-kilowatt engine-driven generator to furnish both alternating and direct current, heating and cooling units to keep the interior within a temperature range of plus or minus ten degrees Fahrenheit, complete facilities for developing oscillograph paper and 16-millimeter film, and means for erecting the antennas from within the body. Storage space for all instruments, spares, and tools for making minor repairs also were included. In addition to the radio telemetery equipment described in the paper, a television telemetering system also was developed to transmit 54 slowly varying items of data (400 cycles per minute). A synchronizing system was developed so that the data recorded from both systems could be correlated for a given instant of time.

Many special instruments were developed in connection with the telemetering project.

Reference

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Negative Resistance Effects in Saturable Reactor Circuits

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Synopsis: Sustained oscillations exhibiting many of the properties of free oscillations are found in saturable reactor circuits under appropriate conditions. The frequencies of such oscillations are classified in relation to that of the driving source as

1. Incommensurable.
2. Subharmonic, and multiples thereof.
3. Harmonic.

General properties of the oscillations classified are elucidated by reference to an analysis of R. V. L. Hartley's, dating back to 1917. In part I a comparatively simple form of analysis is used to demonstrate the development of negative resistance by what is essentially a modulation process. Applications are then made to the three classes of oscillations listed. Some of the simplifying assumptions of part I are removed in part II to approximate practical conditions more closely for the evaluation of a specific product, the third subharmonic. Quantities evaluated over wide ranges of the variables involved include the impedance of the nonlinear coil to the subharmonic, the subharmonic amplitude, its stability, and the conditions required for starting and maintaining the oscillation. Experimental data are found in general agreement with the results of computation.

Part I

THE GENERATION of subharmonics in saturable reactors has had just one commercial application[1] so far as the authors know—the production of 20-cycle telephone ringing current from 60-cycle power. Interest in the phenomenon is far more widespread than the comparatively small field of application would suggest however, as subharmonics may be encountered whenever an attempt is made to tune circuits which include saturable reactors.[2] In fact, subharmonics represent but one of the several classes of sustained oscillations exhibiting similar performance features which may be generated in such circumstances. These classes are arranged according to the ratio of the oscillation frequency to the frequency of the driving a-c source as

1. Incommensurable.
2. Subharmonic, and multiples thereof.
3. Harmonic, both even and odd.

Their performance features differ sharply from those of the more familiar coexistent forced oscillations which include the fundamental and its odd harmonics, assuming the driving source sinusoidal and the nonlinear core unpolarized. To list some of these differences, the oscillations classified can be generated only where the impressed electromotive force falls within a definite range of amplitudes and where the load impedance, too, is located within a restricted region. While the amplitude of any of these oscillations varies continuously with impressed electromotive force or load impedance over a certain range, it falls suddenly and irreversibly to zero when the range is exceeded. Thus the voltage of the generated oscillation goes to zero when the load circuit is opened.

These oscillations have characteristics similar to those of free oscillations in vacuum tube circuits, for example, and

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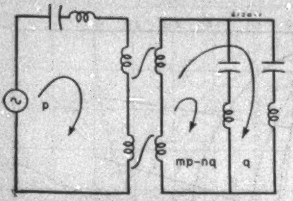

Figure 1. Circuit used by Heegner for developing self-oscillations

The two nonlinear coils are oppositely poled. Resonances are provided at the supply frequency $p/2\pi$, and at the two oscillation frequencies $(mp-nq)/2\pi$ and $q/2\pi$

have been referred to in the literature as free[3] and as self-excited[4] or self oscillations. In any one of the three classes, however, the frequency is related to that of the driving source, a property commonly thought of as an attribute of forced oscillations. Since the oscillations classified therefore have aspects associated with both free and forced oscillations, it might be preferable to designate them by a term indicative of that fact, such as free-forced, combined, composite, or quasi-free. Conforming to previous usage we shall refer to them as free, keeping the reservation just noted in mind.

To provide a physical picture of the way in which free oscillations can be generated in saturable reactor circuits, as well as to exhibit the relationship between the three classes of free oscillations, we shall use the results of a circuit analysis worked out by R. V. L. Hartley[5] in 1917. For this purpose we consider a circuit consisting of a nonlinear coil in series with an external impedance which is assigned suitable values at significant frequencies. The input is supplied by two a-c generators which, for convenient reference, are designated carrier and signal, the term signal being used arbitrarily for the lower of the two input frequencies. With the two input currents flowing through the coil, a whole series of modulation product voltages is generated. The external impedance then is assumed to permit just two of the generated components to flow, the impedance being taken as infinite for the other products.

Solution of the circuit equations permits us to study the effective impedance of the nonlinear coil to the signal. It is found that the flow of certain products introduces a positive resistance, but the flow of certain other products introduces a negative resistance into the signal path. These introduced resistances arise through the intermodulation of the several current components flowing through the coil. The appearance of a negative resistance opens possibilities for the existence of free oscillations. For if the negative resistance can be made large enough to annul the positive circuit resistance at any frequency, and if the circuit reactance can be tuned out at that frequency, the necessary requirements for oscillation are fulfilled.

Because it is fundamental to the operation of magnetic self-oscillating circuits, a brief derivation of the negative resistance (Hartley) effect will be outlined in its simplest form. The results will be compared with corresponding modulation effects in nonlinear resistance circuits. Applications then follow to explain how the three groups of free oscillations can be made to arise. Following this general discussion in part I, the production of a typical subharmonic will be treated quantitatively in part II.

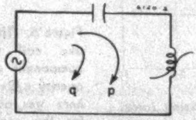

Figure 2. Circuit used by Fallou for frequency reduction

A single resonance is provided for the self-oscillation

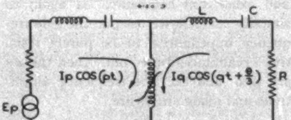

Figure 3. Circuit used for subharmonic analysis with resonances provided at input and subharmonic frequencies

All other current components suppressed

The Hartley Effect

Perhaps the simplest equation we can use to exhibit the relation between flux density and magnetizing force in the saturable reactor is

$$B=a'_1H+a'_3H^3 \qquad (1)$$

This relation exhibits saturation when the nonlinear term is negative, and when the maximum excursion of the magnetizing force is limited to regions in which the slope of equation 1 is positive. With the magnetizing current and magnetizing force linearly related, the voltage developed across the coil may be written

$$v=KB=Ki(a_1+a_3i^2) \qquad (2)$$

The coil is connected in series with a linear impedance and with two independent sine wave generators, their frequencies being $p/2\pi$ and $q/2\pi$, the carrier and signal, respectively. Because of the nonlinearity exhibited by equation 2, new frequencies are generated—in general there are many such components. The frequency of any component may be expressed as $(mp\pm nq)/2\pi$, m and n being integers. Its order is the sum of absolute values of the frequency coefficients $m+n$. This sum will be restricted to odd integers since equation 1 is an odd function and the core is unpolarized. For present purposes there are two nonlinear components in which we are interested primarily. These are the two third-order* side frequencies $(p\pm2q)/2\pi$. To reduce the work to its essentials, we assume networks connected in circuit which permit the flow of just the four components—two impressed by external sources and two gen-

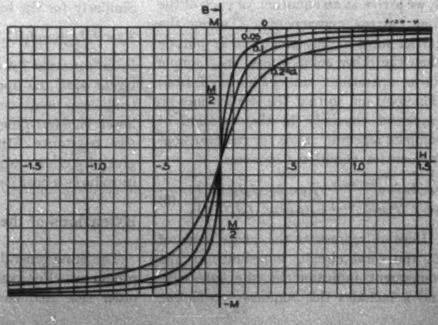

Figure 4. Plot of the equation $B=(2M/\pi)\tan^{-1}(H/a)$ with a, as parameter

This parameter is equal to the magnetizing force at which the flux density reaches half its saturation value M

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* Two other third-order side frequencies ($m=2$ $n=1$) are included in the products which meet a high circuit impedance and are therefore assumed to produce negligibly small current flow.

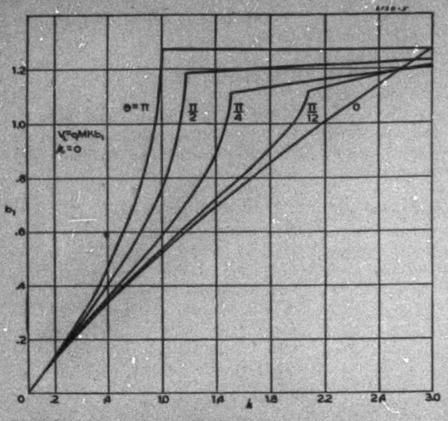

Figure 5. Here a two-component wave of magnetizing force. $H=P\cos 3\,qt+Q\cos(qt+\theta/3)$ is assumed to act on the magnetic characteristic ($a=0$) illustrated in Figure 4

The inductive coil voltage component of frequency $q/2\pi$ is plotted as a function of $k=Q/P$ for several values of θ, $b_1(\theta)=b_1(-\theta)$

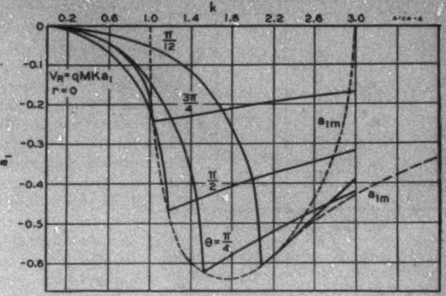

Figure 6. The resistive coil voltage component of frequency $q/2\pi$ shown here was obtained for the conditions specified in Figure 5

$a_1(\theta)=-a_1(-\theta)$,
$a_1=0$ for $\theta=0, \pi$

erated by nonlinearity. Accordingly

$$i=I_p\cos pt+I_q\cos qt+I_+\cos[(p+2q)t+\theta_+]+I_-\cos[(p-2q)t+\theta_-] \quad (3)$$

Here the phase angles $\theta_\pm$ permit necessary phase displacements with respect to the input waves.

The circuit equation may be expressed as

$$\Sigma e_\gamma=\Sigma_\gamma Z_\gamma'i+v \quad (4)$$

where the subscript γ is used to indicate values effective to a particular component p, q, or $p\pm 2q$ in the present case. Σe_γ refers to the externally impressed electromotive forces, v refers as before to the voltage components across the nonlinear coil given by equation 2, and $Z_\gamma'i$ refers to the potential drops across the linear impedance.

Putting equation 3 in equations 2 and 4, we arrive at an equation for each of the four current components. Taking the carrier substantially greater in amplitude than the other components, only simple linear operations are encountered from this point on, so that the results may be expressed concisely in the notation of complex algebra. Thus for the two side-frequencies, equation 4 becomes

$$\left.\begin{aligned}0&=Z_+I_+e^{j(\theta_++\varphi_+)}+j(p+2q)\kappa I_pI_q^2\\0&=Z_-I_-e^{j(\theta_-+\varphi_-)}+j(p-2q)\kappa I_pI_q^2\end{aligned}\right\} \quad (5)$$

Here φ_γ represents the phase angle of Z_γ the total circuit impedance to the component indicated by the subscript, and κ includes the nonlinear coefficient and other coil constants. If again to simplify matters we assume the side frequency impedances to be purely resistive—reactances tuned out—then the φ's are zero, and the side frequency amplitudes and phase angles are

$$\left.\begin{aligned}&\theta_+=\theta_-=-\pi/2\\&I_+=(p+2q)\frac{\kappa I_pI_q^2}{R_+}\\&I_-=(p-2q)\frac{\kappa I_pI_q^2}{R_-}\end{aligned}\right\} \quad (6)$$

These relations permit evaluation of the impedance reaction into the fundamental or input paths. For the lower of the two impressed frequencies, after substituting equation 6 in equations 3 and 4, and again assuming reactances tuned out,

$$E_q=R_qI_q+q4\kappa I_pI_q(I_+-I_-) \quad (7)$$

Similarly for the higher input frequency,

$$E_p=R_pI_p+p\kappa I_q^2(I_+-I_-) \quad (8)$$

From equation 7 it is apparent that in the signal path the upper side-frequency introduces a positive resistance, whereas the lower side-frequency introduces a negative resistance. In the carrier path as given by equation 8, each side-frequency introduces a positive resistance.†

Extensions of the Analysis

The same general result holds for other products. For example, the other third order side frequencies $2p+q$ and $2p-q$ produce positive and negative resistance effects in much the same way as the components $p+2q$ and $p-2q$, respectively. Since the same type of result is found with side frequencies of any order,* the statement following equation 8 can be generalized as follows:

The flow of a lower side-frequency of any order introduces negative resistance into the signal path and positive resistance into the carrier path; the flow of upper side-frequency of any order introduces positive resistance into both paths.

Inasmuch as the term negative resistance is used to describe a source of energy, another way to express this result is to say that the flow of lower side-frequency in a magnetic modulator circuit causes energy to be abstracted from the carrier and delivered to the signal; flow of upper side-frequency abstracts energy from both inputs. Taking into account the transfer of energy from one path to another, it is a straightforward matter to show that energy is conserved.

For subsequent use we can generalize the quantitative results of equations 6 and 7 to fit any order of modulation. Confining attention to the lower side-frequency throughout, equation 6 shows the amplitude of the third order product $(p-2q)$ to be proportional to $I_pI_q^2$. The product of order $m+n$

$$s=mp-nq$$

has an amplitude I_- proportional to $I_p^mI_q^n$, the conclusion following directly from a power series expansion. Equation 7 includes a q voltage component produced by the modulation of the third order product $(p-2q)$ with the two fundamentals to form the product

$$|(p-2q)-p+q|=q$$

This is again a third order process, in-

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† Analogous effects appear in variable capacitor circuits.[6,7]

* Hartley's original analysis dealt with second order side-frequencies, the nonlinear core being polarized.

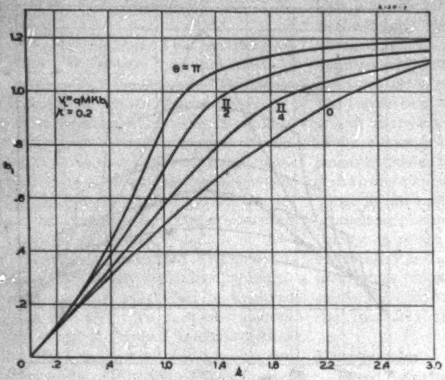

Figure 7. The inductive coil voltage component of frequency $q/2\pi$ corresponds to the condition of Figure 5 except that here $r = a/P = 0.2$, $b_1(\theta) = b_1(-\theta)$

volving three components this time, the amplitude of the reaction term being proportional to $(I_pI_q^2)(I_p)(I_q)$. In general, to get the required q reaction from a modulation involving s, we have to subtract $s-q$, or

$$|(mp-nq)-mp+(n-1)q|=q$$

The amplitude of this product, following the rule previously set forth, is proportional to

$$(I_-)(I_p)^m(I_q)^{n-1}$$

The order of this process is given by the sum of the exponents, 1, m, $n-1$, or $m+n$.

Comparison of Nonlinear Reactive and Resistive Circuits

It is interesting to compare these results with analogous ones found in nonlinear resistance circuits such as diode and copper oxide modulators. Carrying out the work in much the same way as for the nonlinear inductance, it is found,[8,9] that both side-frequencies introduce negative resistance terms into the signal path. Sustained free oscillations never are observed in these circuits, nor are they observed in any circuit in which the slope of the current–voltage characteristic of the nonlinear resistance is always positive. Evidently this differs from the findings of the nonlinear reactance case, and the reason can be stated simply. In the nonlinear reactance case, the negative resistance is produced by a reactance variation; positive resistance components resulting from losses are incidental and comparatively small so that the total circuit resistance can be made negative. Further, the average reactance can be tuned out. In the other case of a resistance variation, the introduced negative resistance is always less than the positive resistance component in any passive element.[8,9] Thus the resistance of the nonlinear element itself is always positive, that of the external circuit adds to it, and the total circuit resistance remains positive.

Free Oscillations

Having established the factors controlling the generation of negative resistance, the next step is to consider applications to some of the consequent oscillation phenomena. First it will be desirable to prevent the flow of the upper side-frequency in order to have as large a negative resistance as possible. Thus we now are considering the current to have just three sinusoidal components in response to the carrier and signal sources. If the external resistance to the side-frequency is now reduced from a large value, the total signal circuit resistance approaches zero.

When this point is reached at a particular value of signal amplitude, the external signal source can be removed[10] and the three components will continue to flow if the circuit reactance is tuned out. Thus, two frequencies are being generated, incommensurable in general with that of the carrier source, the oscillations being of class 1. If, however, the tuning is set so as to make q and p commensurable, the oscillations are of class 2 or of class 3. An interesting point to be noted here is that even harmonics and subharmonics are generated by a purely odd order process,[11] no external polarization being used. Another point is that we may regard the q component as the lower side-frequency in the modulation of p and s, and, similarly, s may be regarded as a lower side-frequency in the modulation of p and q; hence negative resistances are developed in both q and s paths. It should be remarked that no precise adjustment of the reactances is necessary as the circuit is self-adjusting to a certain extent, with the coil impedance depending upon current amplitudes and phases.

It is possible to have negative resistance and oscillation in a circuit in which only two components flow. The process is the same as that outlined previously when we consider that here the signal and side-frequency components have merged to produce an oscillation of class 2 or class 3. Explicitly

$$mp-nq=\pm q$$

or

$$q=mp/(n\pm 1) \qquad (9)$$

Whether free oscillations in any case start from rest or from a finite threshold amplitude can be determined by inspecting the negative resistance term for the q component. This term from the foregoing discussion involves the signal amplitude in the form I_q^{n-1}. Evidently there is just one positive integral value of n (unity) which makes this term independent of I_q and which in turn permits oscillations to start from zero amplitude. Otherwise, with the negative resistance depending upon some positive power of the amplitude, a threshold always will exist below

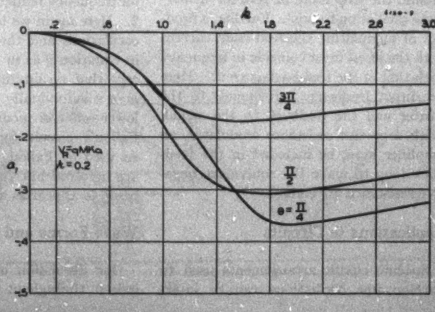

Figure 8. The resistive coil voltage component of frequency $q/2\pi$ plotted here is a companion piece to the inductive component of Figure 7

$a_1(-\theta) = -a_1(\theta)$,
$a_1 = 0$ for $\theta = 0, \pi$

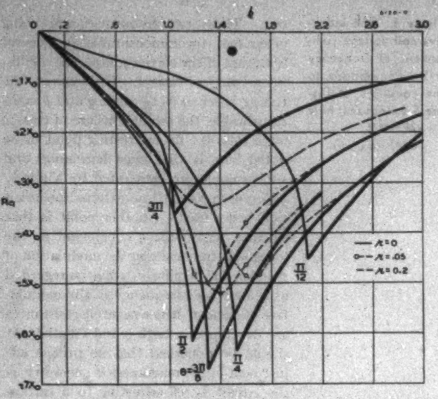

Figure 9. For comparative purposes the resistance of a nonlinear coil at the frequency $q/2\pi$ has been computed for three values of $r=a/P$

The quantity X_0, in terms of which the resistance is expressed, is the coil reactance to the q component when Q is small

$Rq=X_0\lambda a_1/k=0$ for $\theta=0, \pi$

which the total circuit resistance is positive so that free oscillations cannot be sustained. Applying this conclusion to the case of merged components, we have from equation 9 that $q=mp/2$. As $m+n$ is odd and n is unity, m must be an even number. Thus only class 3 oscillations are self-starting where only a single free oscillation flows. But where two individual free oscillations flow, only those oscillations of classes 1 and 2 are self-starting which have the angular velocities $mp-q$ and q.

A vacuum tube circuit yielding results corresponding to equation 9 is the regenerative (resistance) modulator.[12] In that circuit the gain around the feed-back loop must equal unity, and the phase shift around the loop must be zero for steady-state oscillation to exist. Further, a product of modulating the feed-back current with the input must coincide in frequency with that of the feed-back current. Here the input frequency corresponds to the carrier and the feedback to the signal. With a passive resistance modulator, an amplifier must be included in the feed-back loop to make the equivalent negative resistance big enough.

Applications to Circuits

Specific circuit arrangements used to develop free oscillations can be interpreted in the light of the preceding analysis.

A saturable reactor circuit used by Heegner[4,13] is shown in Figure 1. There two distinct nonlinear coils are provided, but the two may be combined in a single somewhat more complex magnetic structure. The use of two nonlinear coils in a balanced circuit permits an easy separation of components. In this arrangement the primary circuit may be tuned to the input frequency. If one of the secondary meshes is tuned to q, the other is tuned to $mp-nq$. With the cores unpolarized, $m+n$ is restricted to odd integers. The balanced coil connection requires m to be even, and so n must be odd. Since, generally speaking, p and q are incommensurable, the oscillations are of class 1.

Another circuit used by Heegner for developing harmonics utilizes just one secondary circuit instead of the two foregoing. This corresponds to merged oscillations in which equation 9 holds. Restricting n to unity yields oscillations of class 3. Allowing n to take higher odd integral values yields, in addition, oscillations of class 2.

The simple circuit used by Fallou[3,14] for frequency reduction is shown in Figure 2. Here again we have a case of merged oscillations and the same relation exists in equation 9 as in Heegner's circuit, except that no limitation is imposed upon m or n individually. As a consequence, a lower order of modulation is required of Fallou's circuit for odd subharmonics. As used by Fallou, oscillations of class 2 are involved but those of class 3 can be made to appear as well.

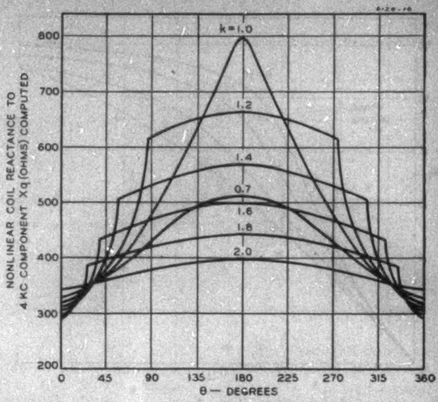

Figure 10. Computed coil reactance to the lower of two impressed frequencies of ratio 1/3, plotted in terms of the phase angle between the two components

The parameter is the ratio of the two component amplitudes. The magnetic characteristic for *a* zero is used here, and the family of curves shown corresponds to the amplitude specified for the higher frequency component

$H=P\cos pt+Q\cos(qt+\theta/3)$

$p/2\pi=12$ kc, $q/2\pi=4$ kc

$P=0.96$ oersted

Wave Forms and Wave Components

Our discussion up to this point and indeed throughout the paper, is based upon the resolution of waves into their Fourier components. This approach is particularly useful where only a few components have to be considered. A different approach, valuable in tracing the minutiae of the starting process, for example, or in dealing with large numbers of components, considers wave forms[15] rather than their components.* In this procedure general principles such as those previously discussed are not discernible, and the analysis is limited to computation of specific numerical cases. The computations are necessarily lengthy as in each case a large number of cycles must be covered to indicate whether or not the oscillation reaches a steady state. When the numbers initially selected result in an oscillation close to the boundary between stable and unstable states, and especially when errors of computations are cumulative, the work is practically interminable. The use of mechanical aids is particularly valuable here. It may be remarked that the great bulk of the literature dealing with wave forms in circuits of the type under discussion is devoted to experimental studes.[3,13]

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* Investigation of this method was carried out by L. W. Hussey and S. O. Rice of these laboratories in 1935.

Part II

The analysis of part I has been designed to present the most important effects in comparatively simple form. Conclusions drawn from this simplified analysis cannot be applied quantitatively to practical cases. Two outstanding reasons are that the *B–H* relation in ferromagnetic materials is not accurately represented by the simple cubic, and that, in general, the significant generated currents are not small compared to those impressed.

Accordingly, we proceed to set up a situation more closely approximating practical conditions which can be analyzed quantitatively.† To facilitate analysis the circuit of Figure 3 is adopted which requires consideration of just two current components, one impressed and the other generated. The generated component, typical of class 2, is a subharmonic one-third the impressed frequency. After the circuit equations are set up in general form, a specific form is introduced for the *B–H* relation. The complexity of the resulting equations makes analytical solution impossible except for limiting cases, and most of the evaluation is graphical. The conclusions are subjected to test in several directions. First the impedance of the nonlinear coil with the two harmonically related frequencies impressed is derived and compared with the results of bridge measurements. Next the self oscillation amplitudes are computed for a variety of circuit conditions and input amplitudes, and compared with measured values. Finally the stability of the oscillations and the conditions required to start them are similarly checked.

Equations for Two-Frequency Magnetization

In this section relations generally valid for two-frequency magnetization are set up under the restrictions previously noted to provide the basis for an analysis of the third subharmonic case. Expressions for the impedance of the nonlinear coil are formulated in terms of Fourier coefficients for the flux density, followed by circuit equations for the oscillation in terms of the same coefficients. The circuit equations permit the oscillation amplitude and phase to vary with time so that stability and starting conditions may be studied.

The magnetizing force acting on the nonlinear core is accordingly

$$
\begin{aligned}
H &= P\cos pt + Q\cos(qt+\varphi)\\
&= P\cos \beta\omega t + Q\cos\left(\alpha\omega t+\frac{\alpha\theta}{\beta}\right) \qquad (10)
\end{aligned}
$$

where α and β are integers and $\alpha/\beta = q/p$ represents the frequency ratio expressed in its lowest terms.

The flux density now can be expressed as a Fourier series having $\omega/2\pi$ for fundamental frequency. Factoring out the saturation value M of the flux density,

$$
B(H) = M\sum_{n=1}^{\infty}\left[a_n \sin n(\omega t+\theta/\beta) + b_n \cos n(\omega t+\theta/\beta)\right] \qquad (11)
$$

The voltage across the coil has a component of frequency $q/2\pi$ equal to

$$
\begin{aligned}
v_q &= K\left[\frac{dB(H)}{dt}\right]_q\\
&= qMK\\
&\quad\times\left[a_\alpha \cos\left(qt+\frac{\alpha\theta}{\beta}\right) - b_\alpha \sin\left(qt+\frac{\alpha\theta}{\beta}\right)\right]
\end{aligned}
$$

and the corresponding q component of current is

$$
i_q = (Q/\Upsilon)\cos\left(qt+\frac{\alpha\theta}{\beta}\right)
$$

where $\alpha\omega = q$, $K = NA\ 10^{-8}$, $\Upsilon = 0.4\ N/d$,

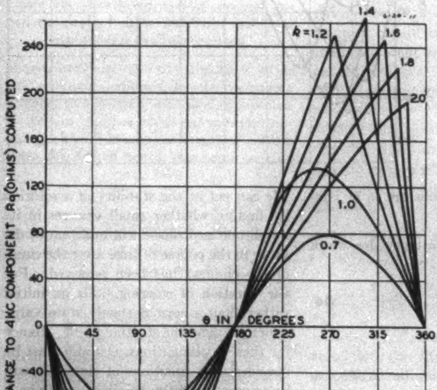


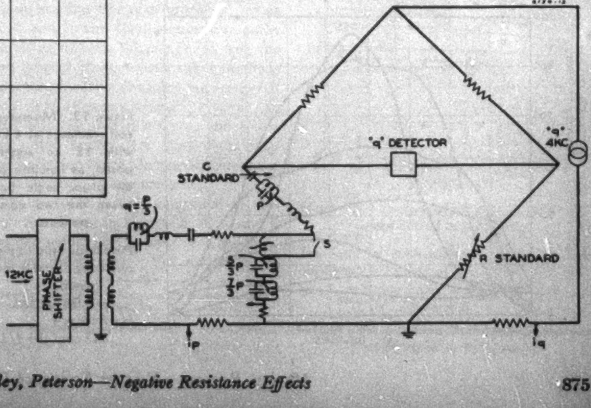
Figure 11. Computed coil resistance to the lower of two input frequencies of ratio 1/3, plotted in terms of the phase angle between the two components

This is a companion piece to Figure 10
$P = 0.96$ oersted

Figure 12 (below). Bridge setup for measuring the impedance components of a nonlinear coil to the lower of two impressed frequencies of ratio 1/3

A phase shifter serves to vary the phase angle between the two input currents

Auxiliary tuned circuits are used to reduce all other current components to a value at least 20 db down on the input currents. Preliminary balance is obtained with switch *S* closed



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† Preliminary work along these lines was done by S. O. Rice in 1934.

and N is the number of conductor turns around the toroidal core of diameter d. From these the impedance of the coil to the q component is defined

$$\begin{aligned} Z(H) &\equiv \frac{v_q}{i_q} \\ &= qMK\tau\left[\frac{a\alpha}{Q}+j\frac{b\alpha}{Q}\right] \\ &= R_q+jX_q \end{aligned} \qquad (12)$$

Defining $Q/P \equiv k$, we have

$$R_q=\frac{X_0\lambda a_\alpha}{k} \qquad (13)$$

$$X_q=\frac{X_0\lambda b_\alpha}{k}$$

where

$$\begin{aligned} X_0 &= qL_{q0} \\ &= \lim_{Q\to 0}\,(X_q) \\ &= \frac{qMK\tau}{\lambda P} \end{aligned} \qquad (14)$$

and λ is specified later in terms of a specific B–H relation. Thus the impedance components of the saturable coil are expressed in terms of L_{q0}, its inductance to a small signal amplitude in the presence of a comparatively large carrier.

Now consider the oscillation q to exist when just one external generator of frequency $p/2\pi$ is connected. Referring to Figure 3, the differential equation for the right-hand mesh, to which the free oscillation is confined, is

$$RS+L\frac{dS}{dt}+\frac{1}{C}\int S\,dt+\tau K\frac{d}{dt}\Big[B(H)\Big]_s=0 \qquad (15)$$

where S is the magnetizing force of the oscillation. Three questions are to be considered:

1. What are the circumstances for which the steady oscillation

$$S=Q\cos\left(qt+\frac{\alpha\theta}{\beta}\right) \qquad (16)$$

is a solution?
2. Is the solution stable?
3. What conditions must be satisfied to start the oscillation?

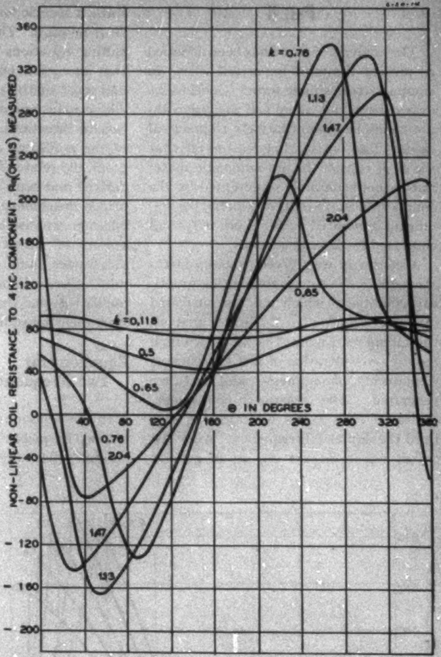

Figure 14. Measured coil resistance at 4 kc with 12 kc superposed, as function of the phase angle between the two components

Amplitude ratio of the two components is the parameter. This is a companion piece to Figure 13
$P=0.96$ oersted

We can get at the stability of a solution by finding whether small changes in its amplitude and phase will build up or die away in the course of time after the cause of the changes has been removed. For the question of starting, it is an initial rather than a normal steady-state value which provides the starting point. Hence for these investigations, Q and θ must be considered as functions of time. It will be assumed that the amplitude and phase vary slowly compared with cos qt. Thus $\frac{dQ}{dt}/pQ$ and $\frac{d\theta}{dt}/q$ are small so that higher order derivatives and products of derivatives can be neglected. It may be seen later that this assumption is consistent with the results. Equations 16, 11, and 14 are inserted into equation 15 and the term $\int(S/C)\,dt$ is changed into an expression involving derivatives through two successive integrations by parts. Then equating coefficients of sine and cosine terms separately to zero, we obtain the two differential equations

$$Rk+\lambda X_0 a_\alpha+\alpha\frac{\lambda X_0 a_\alpha}{\beta q}\frac{d\theta}{dt}+\frac{X_1}{q}\frac{dk}{dt}+\frac{\lambda X_0}{q}\left(\frac{\partial b_\alpha}{\partial k}\frac{dk}{dt}+\frac{\partial b_\alpha}{\partial \theta}\frac{d\theta}{dt}\right)=0 \qquad (17)$$

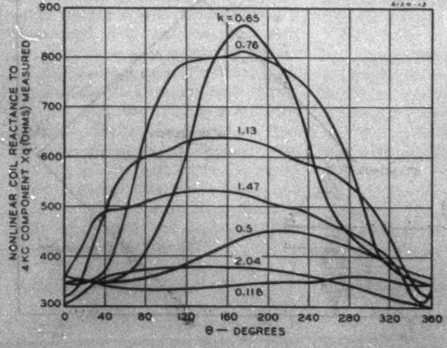

Figure 13. Measured coil reactance at 4 kc with 12 kc superposed, as function of the phase angle between the two components

Amplitude ratio of the two components is the parameter
$P=0.96$ oersted
$H=P\cos pt+Q\cos(qt+\theta/3)$

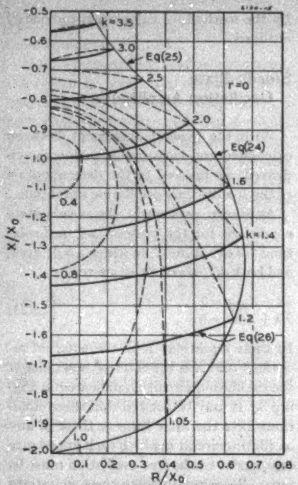

Figure 15. Computed map of external impedance region within which the one-third subharmonic can be generated

For any particular ratio of subharmonic to input amplitude indicated by the parametric numbers, two solutions are found represented by the two branches which join at the boundary. The branch heavily lined represents stable states, the dashed branch represents unstable ones; $r=0$

$$Xk+\lambda X_0 b_\alpha+\alpha\frac{\lambda X_0 b_\alpha}{\beta q}\frac{d\theta}{dt}+\frac{\alpha k}{\beta q}X_1\frac{d\theta}{dt}-\frac{\lambda X_0}{q}\left(\frac{\partial a_\alpha}{\partial k}\frac{dk}{dt}+\frac{\partial a_\alpha}{\partial\theta}\frac{d\theta}{dt}\right)=0$$

Here R and $X=qL-1/qC$ represent the resistance and reactance of the external circuit and $X_1=qL+1/qC$. Equations 17 are simultaneous first order differential equations in Q and θ because higher order derivatives and products of derivatives are negligibly small under the conditions mentioned before. Although we are not able to solve these nonlinear equations completely we can extract from them enough information for our needs.

Steady-state solutions are those for which amplitude and phase remain invariable with time so that dk/dt and $d\theta/dt$ are zero in equation 17. They are to be considered in the next section before taking up stability and starting considerations which require the full equations.

Fourier Coefficients for the Third Subharmonic

We now proceed to calculate the Fourier coefficients b_α and a_α required in the foregoing equation from a particular form of B–H characteristic. The impedance of the saturating coil and the conditions for oscillation then may be determined. As our attention is to be centered upon the third subharmonic representative of class 2 oscillations, we have $\alpha=1$ and $\beta=3$ in equations 10 to 17.

A satisfactory approximation to the form of the B–H relation on each branch of a symmetrical hysteresis loop is given by

$$B(H)=\frac{2M}{\pi}\tan^{-1}\frac{H}{a}\tag{18}$$

M being the saturation value of flux density and a a scale factor for the magnetizing force.* This relation is shown in Figure 4 with a as parameter. It may be noted that in the limit for a zero, the assumed B–H relation approximates an ideal limiter. An approximation sufficiently close for our purposes is to use the same single-valued B–H characteristic for both branches of the hysteresis loop, provided that a supplementary resistance is included in the oscillation circuit to account for hysteresis and eddy losses. This resistance will be treated conveniently as part of the external circuit.

The coefficients a_1 and b_1 of the Fourier series for $B(H)$ are expressed by the usual definite integrals.** When $k\ll 1$, analytical expressions can be derived for the coefficients. These are

$$b_1=\frac{2}{\pi}\frac{k}{\sqrt{1+r^2}}\left[1-\frac{k\cos\theta}{4(1+r^2)}+\dots\right]$$

$$a_1=\frac{2}{\pi}\frac{k}{\sqrt{1+r^2}}\left[-\frac{k\sin\theta}{4(1+r^2)}+\dots\right]\tag{19}$$

where $r=a/P$, and $k\ll 1$. When k is not restricted, the values of b_1 and a_1 for the one-third subharmonic are obtained directly from the definite integrals by computing and plotting values of the integrand as a function of qt for several values of k and θ, and then planimetering the resulting curves. When a or r is zero, the arc tangent characteristic takes the form of a step function changing between $-M$ and $+M$ at zero magnetizing force. In this case the integrals for a_1 and b_1 are computed easily after the values of qt are found for which the magnetizing wave H in equation 10 becomes zero.†

Values of a_1 and b_1 computed in this way are shown in Figures 5, 6, 7, and 8 for two values of r, namely zero and two-tenths, respectively. For accuracy we need to consider an extended range of values of r as the input amplitude P changes. This would involve much labor. The experimental work to be described later centers about a value of r in the neighborhood of 0.05. Examination of significant quantities in this region indicates their rate of variation with r to be fairly small however. This permits us to base our work for the most part on the limiting case of r zero, although on occasion effects introduced by other values of r will be brought in. Further evidence on the consequence of varying r is given in Figure 9, which shows how the derived values R_q/X_0 of equation 13 vary as a function of k, with r and θ as parameters. Three values of r are included: 0, 0.05, and 0.2.

In carrying through the work for $r=0$ as outlined in the appendix, it was found possible to obtain equations for a_1 and b_1 at the salient points of the curves shown in Figures 5 and 6

$$\left.\begin{aligned}a_{1m}&=-\frac{1}{\pi k}[(9-k^2)(k^2-1)]^{1/2}\\ b_{1m}&=\frac{1}{\pi k}(k^2+3)\end{aligned}\right\}\quad k<3\tag{20}$$

A plot of a_{1m} is shown dotted in Figure 6. This is the locus of maximum values of a_1 valid for k less than $\sqrt{5}$. For larger values of k, maximum values of a_1 are represented by the dashed line of the figure for which

$$a_{1M}=-4/\pi k$$

$$b_{1M}=\frac{4\sqrt{(k^2-1)}}{\pi k}\tag{21}$$

Equations 20 and 21 will be found useful in establishing the boundary between stable and unstable states.

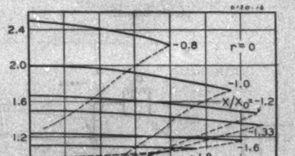

Figure 16. Sections of the map in Figure 15 taken normal to the X-axis

The ratio of subharmonic to input current amplitudes is plotted as a function of the external resistance, with external reactance as parameter

Heavy lines represent stable states, dashed lines represent unstable ones

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* Incidentally, a is a function of frequency when eddy currents are appreciable.

** See Appendix I, equation 31.

† Appendix I.

Coil Impedance to the Oscillation

In this section the values of b_1 and a_1 just found are used to evaluate the resistance and reactance of the coil effective at the signal frequency. The results will be compared with corresponding measured values to provide an estimate of the validity of the approximations used up to this point.

The small signal values of b_1 given by equation 19 may be used to evaluate λ in equation 14

$$\lambda = \frac{\pi}{2}\sqrt{1+r^2} \qquad (22)$$

This is nearly independent of the fundamental input P so long as $P^2 \gg a^2$. Thus the effective inductance to a small signal, by equation 14,* is inversely proportional to the carrier amplitude

$$L_{q0} = \frac{2\tau KM}{P\pi\sqrt{1+r^2}}$$

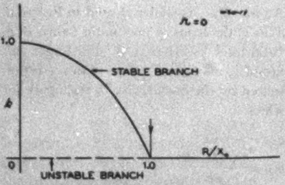

Figure 17. Amplitude curve for a typical class 3 product showing unstable branch lying along the X-axis so that at any external resistance less than the critical, the oscillation can start from rest and grow until it reaches the stable branch

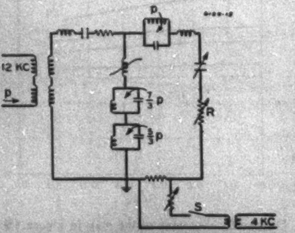

Figure 18. Experimental setup for subharmonic production with resonances provided in input and subharmonic paths

Trap circuits are included to keep the input frequency out of the subharmonic path as well as to suppress unwanted components

S is a starting switch

From the curves of Figures 5 and 6 we may compute the resistance and reactance components given by equation 13. These are plotted in Figures 10 and 11 in terms of the phase angle θ between the two components.

It appears that the effective resistance is negative when the phase angle lies in the first and second quadrants. Inspection of the curves, as well as of the limiting analytic forms, shows that R_q is zero for any order of subharmonic when the oscillation amplitude vanishes, and this is the case for merged oscillations of class 1 or class 2.

Comparison of Calculated and Measured Coil Impedances

Bridge measurements of the impedance components (R_q, X_q) of a saturable coil** were made for comparison with the values calculated previously. Two magnetizing components were supplied by two external generators which were derived from a single source so that the input frequencies of 4 kc and 12 kc were accurately in the 1 to 3 ratio. The 12-kc source, corresponding to the $3q$ component, was introduced through separating tuned circuits into that arm of the bridge which included the saturable coil. The 4-kc source, corresponding to the q component for which the bridge was to be balanced, was introduced at the normal bridge input. Other current components through the coil were reduced to low values by action of the tuned circuits shown in Figure 12. The amplitudes of the two components were individually adjustable, and the phase angle between them was set at will by means of a variable phase shifter. The results are shown in Figures 13 and 14 where R_q and X_q are plotted against θ for various values of k. Comparing them with the computed values of Figures 10 and 11 for r zero shows general agreement in form. The residual resistance resulting from hysteresis, eddy, and copper losses, which may be called R_{q0}, is discernible in the experimental curve for k small as something in the neighborhood of 80 ohms. Supplementary calculation of the coil loss resulting from hysteresis using the arc tangent form for the *B–H* relation and a suitable value for the carrier amplitude yields the value 60 ohms.

Thus measurements show that under certain conditions the saturating coil develops a negative resistance, which is a function of the two magnetizing current amplitudes and of the phase angle between them. This result and in fact the general course of variation of the coil impedance can be approximated roughly by the methods of calculation previously outlined.

Steady-State Oscillation Amplitudes

With the general agreement between calculated and measured impedances evidenced, we proceed to the determination of the steady-state oscillation amplitude.

As shown previously the steady-state conditions for the present case are solutions† of equation 17, in which $\alpha = 1$, $\beta = 3$, and both dQ/dt and $d\theta/dt$ are zero

$$\left.\begin{aligned} Rk + X_0\lambda a_1 &= 0 \\ Xk + X_0\lambda b_1 &= 0 \end{aligned}\right\} \qquad (23)$$

In these equations R is the external resistance and X is the external reactance of the oscillation circuit as indicated in Figure 3. It may be noticed that these equations have the same form as those of any oscillating circuit in which each impedance component to the free oscillation must be zero. We want to solve these equations simultaneously for the oscillation amplitude Q in terms of R and X. This may be done graphically by plotting against each other the values of R/X_0 and X/X_0 satisfying the equations for a particular value of Q or k. When this is done in a number of cases, there results the family of curves shown in Figure 15. For $k < \sqrt{5}$ solutions are found in a region bounded by the circle

$$(R/X_0)^2 + (X/X_0 + 4/3)^2 = (2/3)^2 \qquad (24)$$

while for $k > \sqrt{5}$ the boundary is given by

$$\left(\frac{R}{X_0} - 1\right)^2 + \left(\frac{X}{X_0}\right)^2 = 1 \qquad (25)$$

Equation 24 is obtained by putting equation 20 into equation 23; and equation 25 similarly by inserting equation 21 in equation 23. From equation 24 the largest value of external resistance R at which oscillations can be sustained is

$$R_m = \frac{2}{3}X_0$$

---

* Another relation valid for k small connects L_{q0} with L_{p0}, the inductance of the coil to the p component

$$\frac{L_{p0}}{L_{q0}} = 2\left[1 + r^2 - r\sqrt{1+r^2}\right]$$

As r approaches zero, the ratio approaches two. Note that the inductance ratio is independent of the frequency ratio q/p. A similar relation exists between the corresponding resistance components of a resistance limiter.

** Coil Data:

Core material molybdenum permalloy
Core weight = 1 gram
Tape thickness = 0.6 mil
Core inside diameter = 1.43 centimeters
Core outside diameter = 1.63 centimeters
Cross-sectional core area = 2.4 square millimeters
Number of turns = 200

† One solution of equation 17 is $Q = 0$.

which requires $X=-(4/3)X_0$. Under these conditions $k=1.34$.

It may be remarked that as r increases, the region in which the subharmonic is sustained shrinks. For $r=0.2$, R_m/X_0 is half that for $r=$zero, and for $r=0.05$, R_m/X_0 is eight-tenths that for $r=$zero. The region also shrinks as the subharmonic order increases. Thus for the fifth subharmonic and $r=$zero, R_m is about $^2/_5 X_0'$, X_0' being the initial coil reactance at the oscillation frequency $(1/5)(p/2\pi)$. This amounts to

$$\frac{2}{5}\times\left(\frac{3}{5}X_0\right)$$

Substituting into equation 23 the values of a_1 and b_1 given in Appendix I, equation 34, the heavily lined branches of Figure 15 are represented by the family of circles

$$(R/X_0)^2+(X/X_0)^2=4/k^2 \qquad (26)$$

Inspection of Figure 15 shows that there are two possible values of k for each pair of values of R and X. The situation may be clarified by taking sections through the family of curves, and plotting k against R/X_0 with X/X_0 constant. The double-valued character of the solution is shown clearly in Figure 16. It will be shown later that points on the upper branch (solid line) represent stable solutions, while those on the lower branch (dashed line) are unstable. In using these curves, the coil loss resistance R_{q0} is to be treated as part of the external resistance R. After further inspection of the family of curves of Figure 15 and the method of sectioning, it appears that the stable solutions of Figure 16 are represented by the circles of equation 26.

The average value of oscillation power delivered to the circuit resistance is

$$W_q=\frac{I_q^2R}{2}$$
$$=I_p^2X_0^2\frac{2R}{R^2+X^2}$$

which includes the power lost in the residual resistance R_{q0}. Of course only those values of R/X_0 and X/X_0 lying within the boundary circles of equations 24 and 25 permit oscillations to exist so that the power equation is valid only within the bounded region. For any given value of reactance X, the maximum power is obtained for that value of resistance which is at the boundary of the stable region. From equation 24 the maximum of these limiting powers occurs for $X/X_0=-0.8$ at the value

$$W_{qm}=I_p^2X_0$$

In the case of even harmonics ($\beta=1$) a wholly similar procedure yields the k versus R curve of Figure 17 (solid line). There it is found that if the circuit resistance is less than the critical value

$$R_m=X_0e^{-\frac{\pi r}{\alpha}}$$

even harmonics can be developed. It is interesting to note that the following relations are roughly valid: for harmonic oscillations, $k<1$, and for subharmonic oscillations, $k>1$.

Comparison of Calculated and Measured Oscillation Amplitudes

Oscillation amplitudes observed in the circuit of Figure 18 were compared with the calculations detailed in the preceding section. The nonlinear coil used for the impedance measurements was used here. Auxiliary tuned circuits were included in circuit so that unwanted current components were at least 20 decibels below the input current. Figure 19 shows typical calculated and measured curves for a constant value of fundamental input current and it is seen that the two are comparable. While these results apply to the one-third harmonic, they are typical of class 2 oscillations in a circuit like that of Figure 18. This is borne out by experimental determinations on higher order subharmonics and their multiples, such as the products one-fifth, one-seventh, and four-fifths the fundamental frequency.

Stability and Starting

To investigate the stability of the preceding steady-state solutions, a procedure ordinarily used in dynamics[17] will be followed. The variables k and θ are given values slightly different from the steady-state solutions and differential equations in these small variations are derived. These differential equations then are studied to find out whether in the course of time the variations die away or build up. If they die away so that k and θ return to the steady-state values under investigation, this steady state is said to be stable; if not, it is unstable. In this way the desired answers are obtained without solving equation 17 explicitly.

Accordingly, k and θ are displaced from their steady-state values k_s and θ_s by Δk and $\Delta\theta$ respectively so that

$$k=k_s+\Delta k$$
$$\theta=\theta_s+\Delta\theta \qquad (27)$$

k_s and θ being obtained from equation 23, that is

$$Rk_s+\lambda X_0a_{1s}=0$$
$$Xk_s+\lambda X_0b_{1s}=0 \qquad (23)$$

In the neighborhood of the steady state values, b_1 and a_1 may be expressed as

$$b_1=b_{1s}+\frac{\partial b_1}{\partial k}\Delta k+\frac{\partial b_1}{\partial \theta}\Delta\theta$$

$$a_1=a_{1s}+\frac{\partial a_1}{\partial k}\Delta k+\frac{\partial a_1}{\partial \theta}\Delta\theta \qquad (28)$$

Putting equations 23, 27, and 28 into equation 17 with $\alpha=1$, $\beta=3$ and neglecting quantities involving higher powers of the variations than the first, two simultaneous differential equations in Δk and

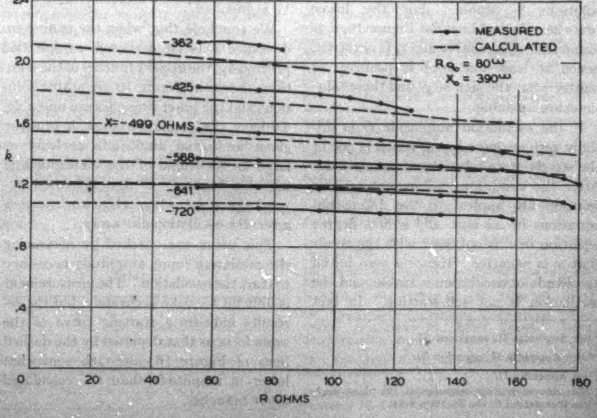

Figure 19. Comparison of measured and computed subharmonic amplitudes as function of external circuit resistance; external circuit reactance as parameter

$\Delta\theta$ are obtained.[*] These equations are linear with constant coefficients with solutions of the form

$$\Delta k = Ae^{ut}$$
$$\Delta\theta = Be^{ut} \qquad (29)$$

Upon inserting these assumed solutions in the differential equations, we get two algebraic equations in A and B, the coefficients of which are functions of u. In order to obtain solutions other than $A=B=0$, it is necessary that the determinant of the coefficients vanish. Setting this determinant equal to zero yields a quadratic equation in u[**]

$$g_2u^2+g_1u+g_0=0 \qquad (30)$$

the solution of which determines u in equation 29. The quantities A and B depend on the initial value of the perturbation, but since their specific values are not needed they will be considered no further.

The character of the roots u of equation 30 determines the stability or lack of it in the steady state. If both roots are negative or have negative real parts, then by equation 29 the variations die away and the steady state is stable. If either of the roots is positive or has a positive real part, then the variations will build up and the steady state is unstable. To investigate the stability of any steady-state solution we must first obtain the corresponding characteristic equation 30 and then determine the nature of its roots.

Consider first the steady-state solutions which are represented by the upper branches (solid lines) of the curves in Figure 16. A study[†] of quadratic equation 30 shows that for all these solutions, the three coefficients g_2, g_1, and g_0 are positive if $X_1/X_0>0.1$; thus both roots u are negative and this group of steady-state solutions is stable. For the lower branches (dashed lines) of Figure 16, it is found that g_2 is positive and g_0 is negative; hence at least one root is positive no matter what the sign of g_1, and these solutions are unstable.

If the oscillation amplitude Q is initially zero we have $k_s=0$, which is one of the steady-state solutions of equation 23. After the conditions accompanying this solution are applied to the differential equations in Δk and $\Delta\theta$,[†] a first degree equation in u is obtained with the result that u is negative. Hence a zero initial amplitude of oscillation is stable, and the oscillation is not self-starting. In fact the oscillation will decrease to zero if k is initially less than the value on the lower branch, and will increase if k is initially greater than this value. In both cases there is a positive u, but the variations are opposite in sign. Another way of expressing these facts is to say that the oscillating circuit has a negative resistance for amplitudes between the stable and unstable values, and a positive resistance outside this range; the oscillation builds up when the over-all resistance is negative. A study of this resistance when R/X_0 is increased beyond the limiting value (0.65 for $X/X_0=-1.2$), shows that it is positive for all values of k, and so an oscillation will die away no matter what the initial amplitude may be. These considerations account for the necessity of providing starting means in this circuit. The same requirement has been established experimentally by Fallou[14] for his single-mesh circuit.

The limitation $X_1/X_0>0.1$ mentioned previously is not restrictive as in any practical case it is necessary for X_1/X_0 to be substantially greater than unity in order that the oscillation may be sinusoidal as assumed in equation 10. From the definitions of X_1 and X, we have

$$X_1/X_0=2qL/X_0-X/X_0$$

so that X_1 increases with the series inductance L. In the experimental work $gL=2X_0$ approximately, hence $X_1/X_0=3$ nearly. When X_1/X_0 is of this magnitude, the roots u of equation 24 are roughly proportional to X_0/X_1. In particular, if $X_1/X_0>3$, then $u<0.1\ q$ and thus $\frac{1}{qk}\frac{dk}{dt}$ and $\frac{1}{q}\frac{d\theta}{dt}$ are small. Therefore the assumption made before in deriving equations 17 is justified.

We conclude that when the tuning inductance L is large enough as specified previously, the upper branches of the solutions shown in Figure 16 are stable solutions and the lower branches are unstable. To have the oscillation start, it must be given an initial amplitude at least as large as the value of the corresponding unstable branch. If R is increased beyond the point where the two branches meet, the oscillation dies away.

This point was checked by measuring the minimum input amplitude necessary to start the oscillation. The measurement is difficult to make accurately, but the results indicate a starting curve of the same form as that depicted by the dashed lines of Figure 16, although somewhat lower in amplitude than the calculated lower branches.

Table I

	$0 \le \theta \le \pi$	
	$k<k_0$	$k>k_0$
$\partial b_1/\partial k$	+	+
$\partial a_1/\partial k$	−	+

For the even harmonic case, a typical k versus R curve is found to look like the one shown in Figure 17. So long as the circuit resistance is less than the critical value, any small disturbance will build up to the upper (stable) branch as the R axis ($k=0$) corresponds to the lower unstable branch. Hence the oscillation is self-starting, as found experimentally.

It may be concluded, therefore, that the general picture of oscillating amplitudes, stabilities, and starting conditions is reproducible by the methods of calculation employed.

It should be emphasized that our discussion has been confined to the particular case in which just one generated current component is permitted to flow. When additional products flow, complications appear because some introduce positive, and others negative resistance effects which may cause the circuit performance to change radically. For example, if with the one-third subharmonic flowing a low impedance path is provided for its fifth harmonic, the oscillations are made self-starting.[15]

Appendix I

The coefficients a_1, b_1 of the Fourier series for $B(H)$ in equation 11 are expressed by the usual integrals after inserting equation 10 in equation 18, that is,

$$b_1=\frac{2}{\pi^2}\times\int_0^{2\pi}\tan^{-1}\left[\frac{\cos 3qt+k\cos(qt+\theta/3)}{r}\right]\times\cos(qt+\theta/3)dqt$$

$$a_1=\frac{2}{\pi^2}\times\int_0^{2\pi}\tan^{-1}\left[\frac{\cos 3qt+k\cos(qt+\theta/3)}{r}\right]\times\sin(qt+\theta/3)dqt \qquad (31)$$

where $r=a/P$. By simple transformations it may be shown that

$$b_1(-\theta)=b_1(\theta)$$
$$a_1(-\theta)=-a_1(\theta)$$

Alternatively we could compute and sum all the modulation products of frequency $q/2\pi$. This is equivalent to expanding equations 31 in Fourier series in θ. The direct method is the more desirable because it gives a complete result, whereas the second usually requires the summation of a large number of terms for a comparable result. When $k \ll 1$ in equation 31, the analytical expressions in equation 19 can be derived for the coefficients by expanding the arc tangent function of the integrand in a power series in $k \cos (qt+\theta/3)$.

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[*] See Appendix II, equations 35.

[**] See Appendix II, equation 36.

[†] See Appendix II.

[‡] For convenience in computation, the phase angle θ was transferred to the $3qt$ term here.

In computing a_1 and b_1 by the direct method when $r=0$, we have instead of equation 31‡

$$b_1=\frac{2}{\pi}\int_0^{\pi} \pm \cos y\, dy$$

$$a_1=\frac{2}{\pi}\int_0^{\pi} \pm \sin y\, dy \qquad (32)$$

The positive sign is used when H in equation 10 is positive and the negative sign when H is negative. In general, there are three values of qt between 0 and π at which $H=0$, designated as y_1, y_2, and y_3. For $k<k_0$, k_0 being a function of θ, the positive sign is used from 0 to y_1 and from y_2 to y_3, and the negative sign from y_1 to y_2 and from y_3 to π. Carrying through the indicated operations, we get

$$b_1=(4/\pi)(\sin y_3+\sin y_1-\sin y_2)$$

$$a_1=(-4/\pi)(\cos y_3+\cos y_1-\cos y_2) \qquad (33)$$

As k approaches k_0, y_3 and y_2 approach each other until at $k=k_0$, $y_3=y_2$ and a_1 and b_1 are at the salient points as shown on the curves. For $k>k_0$ there is only one value y_1 at which H is zero. In this case as at $k=k_0$,

$$b_1=(4/\pi)(\sin y_1)$$

$$a_1=(-4/\pi)(\cos y_1) \qquad (34)$$

To facilitate calculation, y_1, y_2, and y_3 were computed and curves were plotted as functions of k and θ.

At the point where $y_3=y_2$ and the salient points of a_1 and b_1 appear, we have $dk/dy=0$ and $dH/dy=0$. Expressing this with the aid of equation 10 and solving simultaneously with $H=0$ enables us to get analytical expressions for $\sin y_1$ and $\cos y_1$ at these points and hence the expressions in equation 20 for a_1 and b_1.

For values of $k>\sqrt{5}$, these salient points are not the maximum values for a_1. The maximum values, a_{1M} in equation 21, were found by setting $da_1/\partial\theta=0$.

Appendix II

The two differential equations in Δk and $\Delta\theta$ mentioned just preceding equation 29 are

$$\left(R+\lambda X_0\frac{\partial a_1}{\partial k}\right)\Delta k+\left(\frac{X_1}{q}+\frac{\lambda X_0}{q}\frac{\partial b_1}{\partial k}\right)\frac{d}{dt}(\Delta k)+$$
$$\lambda X_0\frac{\partial a_1}{\partial\theta}\Delta\theta+\left(\frac{\lambda X_0}{q}\frac{\partial b_1}{\partial\theta}-\frac{1}{3}\frac{k_sR}{q}\right)\times$$
$$\frac{d}{dt}(\Delta\theta)=0 \qquad (35)$$

$$\left(X+\lambda X_0\frac{\partial b_1}{\partial k}\right)\Delta k-$$
$$\frac{\lambda X_0}{q}\frac{\partial a_1}{\partial k}\frac{d}{dt}(\Delta k)+\lambda X_0\frac{\partial b_1}{\partial\theta}\Delta\theta+$$
$$\left[\frac{k_s}{3_q}(X_1-X)-\frac{\lambda X_0}{q}\frac{\partial a_1}{\partial\theta}\right]\frac{d}{dt}(\Delta\theta)=0$$

Explicitly equation 30 is

$$\frac{u^2}{q^2}\left[\frac{k_s}{3}\left(\frac{X_1}{X_0}-\frac{X}{X_0}\right)\left(\frac{X_1}{X_0}+\frac{\lambda\partial b_1}{\partial k}\right)-\right.$$
$$\frac{\lambda X_1}{X_0}\frac{\partial a_1}{\partial\theta}-\frac{\lambda k_s}{3}\frac{R}{X_0}\frac{\partial a_1}{\partial k}+$$
$$\left.\lambda^2\left(\frac{\partial b_1}{\partial\theta}\frac{\partial a_1}{\partial k}-\frac{\partial b_1}{\partial k}\frac{\partial a_1}{\partial\theta}\right)\right]+$$
$$\frac{u}{q}\left[\frac{k_s}{3}\frac{R}{X_0}\frac{X_1}{X_0}+\lambda\left(\frac{X_1}{X_0}-\frac{X}{X_0}\right)\times\right.$$
$$\left.\left(\frac{\partial b_1}{\partial\theta}+\frac{k_s}{3}\frac{\partial a_1}{\partial k}\right)+\frac{\lambda R}{X_0}\left(\frac{k_s}{3}\frac{\partial b_1}{\partial k}-\frac{\partial a_1}{\partial\theta}\right)\right]+$$
$$\left[\frac{\lambda R}{X_0}\frac{\partial b_1}{\partial\theta}-\frac{\lambda X}{X_0}\frac{\partial a_1}{\partial\theta}+\right.$$
$$\left.\lambda^2\left(\frac{\partial b_1}{\partial\theta}\frac{\partial a_1}{\partial k}-\frac{\partial b_1}{\partial k}\frac{\partial a_1}{\partial\theta}\right)\right]=0 \qquad (36)$$

As shown here, stability considerations require determination of the signs of g_2, g_1, and g_0, which are the coefficients in equation 36. A brief outline of this determination and some of the intermediate results are given here.

The partial derivatives of a_1 and b_1 were found graphically except in the region near $k=k_0$ where they are large for $k<k_0$. Supplementary expressions for a_1 and b_1 in the neighborhood of $k=k_0$ were worked out and derivatives computed. The Jacobian

$$\frac{\partial b_1}{\partial\theta}\frac{\partial a_1}{\partial k}-\frac{\partial b_1}{\partial k}\frac{\partial a_1}{\partial\theta}$$

which appears in both g_0 and g_2, is zero for $k>k_0$ as may be verified by means of equation 34; for other values of k, it is small and negative. The term X/X_0 is always negative because b_1 is positive and the two are connected as shown in equation 23.

The signs of the partial derivatives are given in Tables I, II, and III.

In Table II θ_0 is the value of θ corresponding to the salient point for a given value of k, and θ_1 is that value for which $\partial a_1/\partial\theta=\partial b_1/\partial\theta=0$.

Applying all the information just mentioned to equation 36, it is found that when $X_1/X_0>0.1$, g_2, g_1, and g_0 have signs as as shown in Table III.

Thus the dividing point between stable and unstable solutions occurs at the point where $\partial a_1/\partial\theta$ changes sign. As seen from the table of derivative signs, this point occurs at the salient point (k_0, θ_0) when $k<\sqrt{5}$ and at $\theta_1[\cos\theta_1=(3-4/k^2)/k]$ when $k>\sqrt{5}$. A distinction is made between the two regions because $\partial a_1/\partial\theta$ changes sign discontinuously at the salient points. Referring to the curves k versus R/X_0, X/X_0 constant, shown in Figure 16, θ increases as k increases and R/X_0 increases as a_1 increases. Hence the upper branches of these steady-state curves are stable and the lower branches are unstable.

Table II

	$k<\sqrt{5}$		$k>\sqrt{5}$		
	$\theta<\theta_0$	$\theta>\theta_0$	$\theta<\theta_0$	$\theta_0<\theta<\theta_1$	$\theta>\theta_1$
$\partial b_1/\partial\theta$	+	+	+	−	+
$\partial a_1/\partial\theta$	−	+	−	−	+

Table III

	$\partial a_1/\partial\theta>0$	$\partial a_1/\partial\theta<0$
g_2	+	+
g_1	+	
g_0	+	−

For the steady state $k_s=0$, we have from equation 19

$$\partial a_1/\partial\theta=\partial a_1/\partial k=\partial b_1/\partial\theta=0$$
$$\partial b_1/\partial k=2\pi$$

Putting these in the differential equations for Δk and $\Delta\theta$, we find

$$u=\frac{-R_q}{X_1+X_0}<0$$

so that the zero amplitude solution is stable.

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Rotating Electric Machine Time Constants at Low Speeds

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THE use of the concept of time constants in the analysis of a-c rotating electric machine transients is well known.[1] It is also well known that the short-circuit time constants of machines running at normal speed are not at all like those of the same machines at rest. The purposes of this paper are to show the continuous nature of the variation of the machine characteristics in the range of speed from zero to rated speed, and to determine the limits of applicability of the full-speed and standstill values and the behavior at low speed.

Rated-Speed Characteristics

In the analysis of transient currents and voltages in induction motors and synchronous motors and generators operating at rated speed, a procedure is more or less well established.[1] In the case of induction motors,* there is a d-c time constant for the decay of the d-c component of current given by the ratio of the transient (short-circuit or standstill) reactance to the stator resistance; there is also an a-c time constant, for the decay of the a-c component of current, given by the ratio of the transient reactance to the rotor resistance. In the case of synchronous machines, there are similarly a d-c (or armature) time constant and two a-c (a subtransient and a transient or field short-circuit) time constants. The two a-c or rotor time constants used most frequently are the direct-axis quantities, the quadrature-axis rotor time constant being considered to have only a secondary effect on short-circuit conditions. The d-c time constant is given by the ratio of the negative-sequence reactance (or the average of the direct- and quadrature-axis subtransient reactances) to the armature resistance; the direct-axis subtransient time constant is given by the ratio of the short-circuit reactance viewed from the direct-axis amortisseur to the (direct-axis) amortisseur resistance; the field short-circuit time constant is given similarly by the ratio of the short-circuit reactance viewed from the field (and neglecting the amortisseur) to the field resistance. One notes in all of these definitions the rather remarkable simplifying feature that each time constant depends on only one resistance; there is apparently no interaction between stator and rotor resistances. The reason for this lies in the relative motion of the rotor and stator. All of these time constants are fundamentally d-c time constants for the circuit in question, and a d-c or slowly decaying current in any winding corresponds to an alternating (or speed frequency) current in a winding which is rotating with respect to the first. The resistance in the winding containing the alternating current is usually negligible compared to the reactance at the frequency in question. The apparent separation of the field and direct-axis amortisseur resistances is qualitatively quite different and arises only because of the usually very large difference in magnitude between these two resistances; it is not a generality and would not be true for an unusually high field resistance or an unusually low amortisseur resistance.

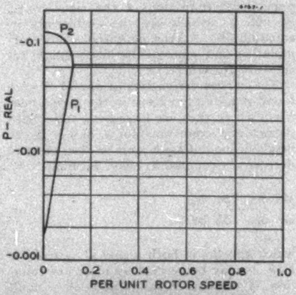

Figure 1. Induction motor—decrement factors as functions of rotor speed

(Real parts of roots of equation 4)

$x_1 = x_2 = 0.08$
$x_m = 3.00$
$r_1 = r_2 = 0.01$

Standstill Characteristics

When the machine is at rest the situation is entirely different and the rated-speed time constants have no meaning. In the case of an induction motor the d-c time constant, like that of a transformer, ordinarily is considered to be given by the ratio of the short-circuit reactance to the sum of the stator and rotor resistances, and thus may be less than half of the full-speed value. Actually there are two d-c time constants, one defined in the preceding sentence, or more nearly by the ratio of the sum of the stator and rotor leakage reactances to the sum of the stator and rotor resistances, the other given approximately by the ratio of the sum of the mutual (magnetizing) reactance plus the paralleled stator and rotor leakage reactances to the paralleled stator and rotor resistances. The first time constant is the only one apparent in an oscillogram of the starting current. The second time constant

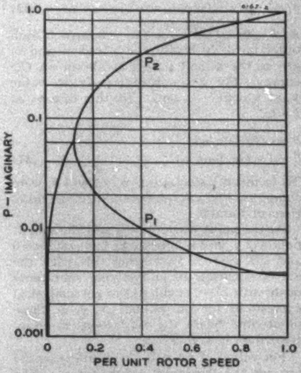

Figure 2. Induction motor—natural frequencies as functions of rotor speed

(Imaginary parts of roots of equation 4)

$x_1 = x_2 = 0.08$
$x_m = 3.00$
$r_1 = r_2 = 0.01$

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* It tacitly is assumed here and in the discussion of synchronous machines that stator currents rather than rotor currents are being computed, and that the machine is either connected to an infinite system or has all three phases short-circuited. Further, we are dealing entirely with calculated values, and the time constants discussed are defined in terms of the calculation required to obtain them. A practical definition must, of course, admit of determination by a not too complicated measurement, and may be incompatible except as an approximation with the abstractions dealt with here. This question is discussed in reference 2.

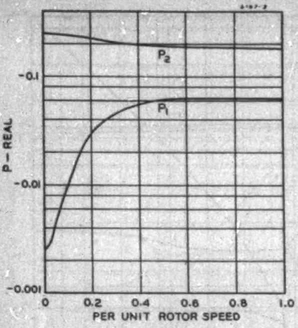


Figure 3. Induction motor—decrement factors as functions of rotor speed

(Real parts of roots of equation 4)

$x_1 = x_2 = 0.08$
$x_m = 3.00$
$r_1 = 0.01$
$r_2 = 0.03$

applies to only a very small component of stator or rotor current, which is however a magnetizing current and corresponds to a rather large flux and torque.[4]

For a synchronous machine, similar but more complex conditions apply. Even the most elementary considerations for a machine with no amortisseur will show that there are two armature (stator) d-c time constants, one for each rotor axis. In the direct axis the time constant is given by the ratio of the transient reactance (or more nearly the sum of the armature and field leakage reactances) to the sum of the armature and field resistances; in the quadrature axis it is given by the ratio of the quadrature-axis reactance to the armature resistance. These values may differ by a factor of two or more to one. There is also, as in case of the induction motor, a direct-axis flux time constant given by the ratio of the sum of the mutual reactance plus the paralleled armature and field leakage reactances to the paralleled field and armature resistances.

Low-Speed Characteristics

The procedures at full speed and at rest are relatively clear and straightforward. However, because the conclusions differ so markedly the question naturally arises as to what happens at very low speeds, and where the transition between the two forms takes place, if there is any such definite speed. In order to show the continuity of the characteristics as functions of speed and to determine the lowest speed at which the full-speed concepts may be used, this paper presents in Figures 1–12 the results of calculations of the various short-circuit time constants at all speeds from zero to full speed for several more or less typical machines.

The figures show decrement factors (reciprocals of the time constants) and natural frequencies calculated from the complete system equations. Figures 1 and 2 are for an induction motor with equal stator and rotor resistances. Figures 3 and 4 are for the same motor except that the rotor resistance is now three times as great as before. These figures show that strictly speaking there is really no such thing as a d-c time constant when the machine is running. The component that may appear in an oscillogram as a decaying direct current has actually a period at full speed ($\omega = 1.0$) of about 250 normal cycles for the case of Figure 2 (a rotor resistance of 0.01), and 100 cycles for the case of Figure 4 (a rotor resistance of 0.03). Similarly, the normal frequency transient component has really only a nearly normal frequency, being necessarily of such a frequency that the sum of the frequencies of the two components equals rotor speed. The real oscillatory nature of the d-c component becomes more pronounced at low rotor speeds, the case of equal stator and rotor resistances (Figure 2) being especially remarkable in that the frequencies of both transient components become exactly equal to one-half rotor speed below a speed of about 12 per cent. The value 12 per cent is seen to be a critical speed below which the frequencies are equal and above which the decrements are equal (Figure 1). Moreover, the decrements in this case are given exactly by the ratio of the transient reactance to the stator (or rotor) resistance, although ordinarily this formula must be regarded as only approximate.

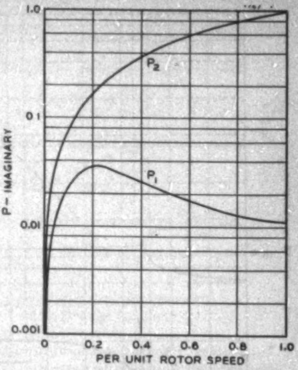


Figure 4. Induction motor—natural frequencies as functions of rotor speed

(Imaginary parts of roots of equation 4)

$x_1 = x_2 = 0.08$
$x_m = 3.00$
$r_1 = 0.01$
$r_2 = 0.03$

The full-speed time constants are given approximately by the relations

$$\text{Stator time constant} = \frac{x_1 + \dfrac{x_2 x_m}{x_2 + x_m}}{r_1} = \frac{x'}{r_1} \qquad (1)$$

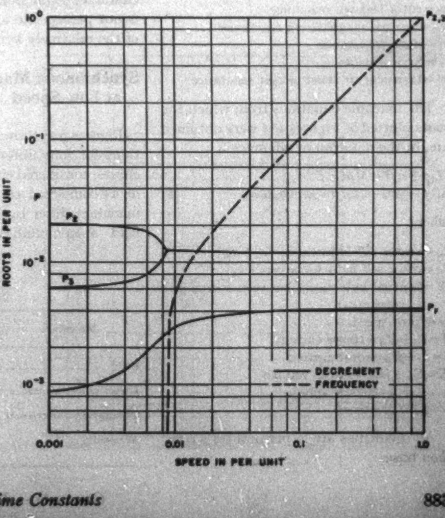


Figure 5. Synchronous machine—decrement factors and natural frequences

(Roots of the characteristic equation 9)

No amortisseur
$x_q = 0.8$
$x_d = 1.2$
$x_d' = 0.2909$
$T_o = 1{,}000$
r_a 0.005

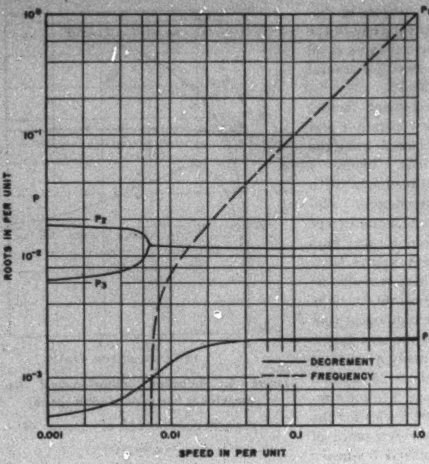


Figure 6. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

No amortisseur
$x_q=0.8$
$x_d=1.2$
$x_d'=0.2909$
$T_o=1,000$
$r_a=0.005$

$$\text{Rotor time constant}=\frac{x_2+\dfrac{x_1x_m}{x_1+x_m}}{r_2} \qquad (2)$$

where

x_1 = stator leakage reactance
x_2 = rotor leakage reactance
x_m = mutual magnetizing reactance
r_1 = stator resistance
r_2 = rotor resistance
x' = transient or short-circuit reactance

The machine equations from which the roots plotted in Figures 1–4 were obtained are, in the notation of reference 3,

$$(L_1p+r_1)I+Mpi=E$$
$$M(p-j\omega)I+[L_2(p-j\omega)+r_2]i=e \qquad (3)$$

where

L_1 = stator self inductance $=x_1+x_m$
L_2 = rotor self inductance $=x_2+x_m$
M = mutual inductance between stator and rotor $=x_m$
ω = rotor speed
$I=I_\alpha+jI_\beta$ = stator current
$i=i_\alpha+ji_\beta$ = rotor current
$E=E_\alpha+jE_\beta$ = stator voltage
$e=e_\alpha+je_\beta$ = rotor voltage
$p=d/dt$ (t = time)

All quantities are in per unit on a common base.

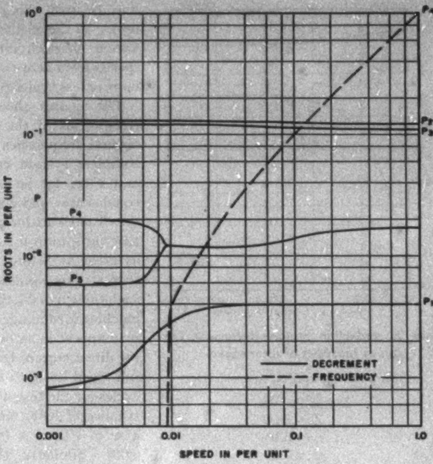


Figure 7. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=0.2$	$x_f=0.1$
$x_{aq}=0.6$	$r_a=0.005$
$x_{ad}=1.0$	$r_{kq}=0.04$
$x_{kq}=0.2$	$r_{kd}=0.02$
$x_{kd}=0.1$	$r_f=0.0011$

The characteristic equation of the coefficients of these performance equations is

$$(L_1p+r_1)(L_2p+r_2)-M^2p^2-j\omega L_2(L_1p+r_1)+j\omega M^2p=0 \qquad (4)$$

Conclusions for Induction Motors

The results shown here indicate that the full-speed decrements are more or less effective from full speed down to rather low speeds (about 10 to 40 per cent). In the low-speed range the more nearly correct values given here must be used for accurate work. The oscillatory character of the nominally d-c component is not perceptible at rated speed because of the relatively very rapid decay.

Synchronous Machines at Low Speed

Figures 5–8 show the results of calculations for four different synchronous machines, considered either as short-circuited or as connected to an infinite bus. The machines differ by having two values of field open-circuit time constant, $T_0=$ 1,000 per unit (Figures 5 and 7) and $T_0=$ 2,000 (Figures 6 and 8) and by being considered with (Figures 7 and 8) and without (Figures 5 and 6) and amortisseur. These figures show that for the synchro-

Table I

	Assumption			
Short-Circuit Time Constant	r_a	r_f	r_{kd}	r_{kq}
Field	0*	r_f	∞	r_{kq}
Direct-axis amortisseur	0*	0	r_{kd}	r_{kq}
Quadrature-axis amortisseur	0*	r_f	r_{kd}	r_{kq}
Armature	r_a	0	0	0

* In these cases, instead of assuming $r_a=0$ it is possible to assume $\omega=\infty$.

Table II. Short-Circuit Time Constants

Machine		Figure 5	Figure 6	Figure 7	Figure 8
Field	Exact	242	485	248	490
	Approximate	242	485	242	485
Direct amortisseur	Exact			7.94	8.0
	Approximate			8.125	8.125
Quadrature amortissuer	Exact			8.75	8.75
	Approximate			8.75	8.75
Armature	Exact	85.3	85.4	58.2	58.4
	Approximate	85.3	85.3	58.0	58.0

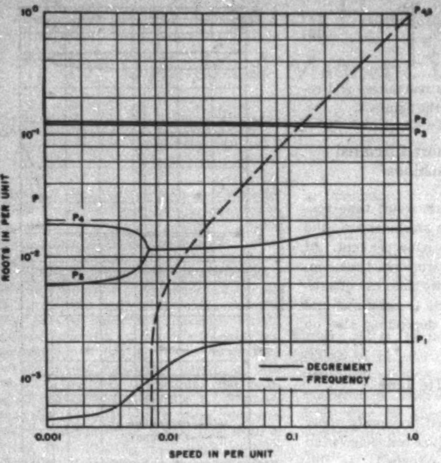


Figure 8. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=0.2$
$x_{aq}=0.6$
$x_{ad}=1.0$
$x_{kq}=0.2$
$x_{kd}=0.1$
$x_f=0.1$
$r_a=0.005$
$r_{kq}=0.04$
$r_{kd}=0.02$
$r_f=0.00055$

nous machines considered, the decrement factors, and thus the time constants, have nearly their full-speed values down to a very low speed (from about one to five per cent). Below about one per cent speed all of the roots become real. It should be noted that, because the machine equations are set up in terms of axes fixed in the rotor, the armature d-c decrements (P_2 and P_3 in Figures 5 and 6, P_4 and P_5 in Figures 7 and 8) are the real parts of the complex roots, and correspond to the speed-frequency oscillations of direct- and quadrature-axis currents.

It is interesting to compare the values of short-circuit time constant found at full speed from the rigorous solution of the characteristic equation with the values found from the usual approximate formulas, which may be written as

$$\text{Field time constant}=\frac{1}{r_f}\left[x_f+\frac{x_l x_{ad}}{x_d}\right] \qquad (5)$$

Direct-axis amortisseur time constant =

$$\frac{1}{r_{kd}}\left\{x_{kd}+\left[\frac{1}{x_f}+\frac{1}{x_{ad}}+\frac{1}{x_l}\right]^{-1}\right\} \qquad (6)$$

Quadrature-axis amortisseur time constant

$$=\frac{1}{r_{kq}}\left[x_{kq}+\frac{x_l x_{aq}}{x_q}\right] \qquad (7)$$

$$\text{Armature time constant}=\frac{2}{r_a}\left[\frac{x_d''x_q''}{x_d''+x_q''}\right] \qquad (8)$$

where

x_l = armature leakage reactance
x_f = field leakage reactance
r_f = field resistance
r_{kd} = direct-axis amortisseur resistance
r_{kq} = quadrature-axis amortisseur resistance
x_{kd} = direct-axis amortisseur leakage reactance
x_{kq} = quadrature-axis amortisseur leakage reactance
r_a = armature resistance
x_{ad} = direct-axis mutual reactance (assumed to be the same for armature, field, and amortisseur)*
$x_d=x_{ad}+x_l$
x_{aq} = quadrature-axis mutual reactance between armature and amortisseur

$x_q=x_{aq}+x_l$

Figure 9. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=1.2$ $x_f=0.1$
$x_{aq}=0.6$ $r_a=0.005$
$x_{ad}=1.0$ $r_{kq}=0.04$
$x_{kq}=0.2$ $r_{kd}=0.02$
$x_{kd}=0.1$ $r_f=0.0011$

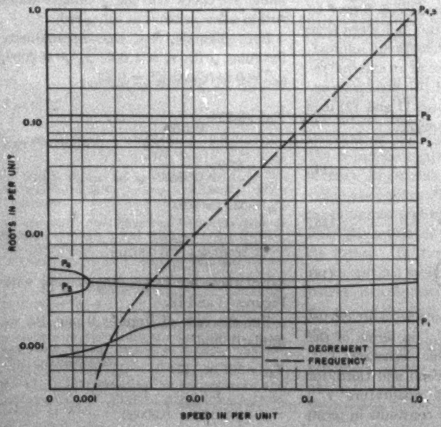
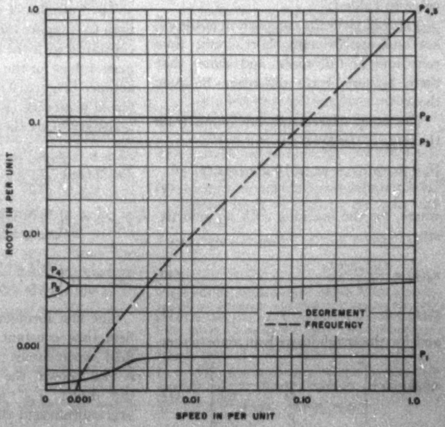


Figure 10. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=1.2$ $x_f=0.1$
$x_{aq}=0.6$ $r_a=0.005$
$x_{ad}=1.0$ $r_{kq}=0.04$
$x_{kq}=0.2$ $r_{kd}=0.02$
$x_{kd}=0.1$ $r_f=0.00055$



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* This may be accomplished by choice of rotor current bases as in definition 4 of reference 5.

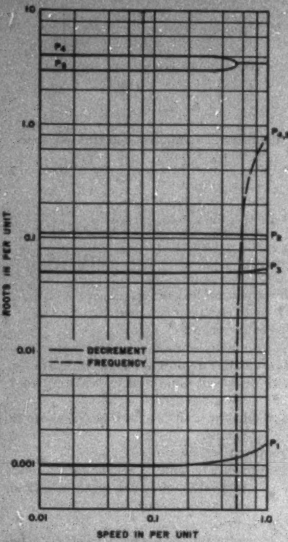

Figure 11. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=0.2$ $x_f=0.1$
$x_{aq}=0.6$ $r_a=1.005$
$x_{ad}=1.0$ $r_{kq}=0.04$
$x_{kq}=0.2$ $r_{kd}=0.02$
$x_{kd}=0.1$ $r_f=0.0011$

All quantities are in per unit on a common base.

These approximate formulas are obtained by making assumptions as indicated in Table I.

Table II gives the comparison of the exact and approximate short-circuit time constants at full speed, and shows that there is very little difference between them.

The exact decrements are found from the roots of the equation

$$[r_a+px_d(p)][r_a+px_q(p)]+\omega^2x_d(p)x_q(p)=0 \qquad (9)$$

where, for the machine with no amortisseur,

$$x_d(p)=\frac{x'_dT_op+x_d}{T_op+1} \qquad (10)$$

$$x_q(p)=x_q \qquad (11)$$

and for the machine with an amortisseur,

$$x_d(p)=x_l+\left[\frac{1}{x_{ad}}+\frac{1}{x_f+\frac{r_f}{p}}+\frac{1}{x_{kd}+\frac{r_{kd}}{p}}\right]^{-1} \qquad (12)$$

$$x_q(p)=x_l+\left[\frac{1}{x_{aq}}+\frac{1}{x_{kq}+\frac{r_{kq}}{p}}\right]^{-1} \qquad (13)$$

Values of the various reactances and resistances are given on the figures.

Conclusions for Short-Circuited Synchronous Machines

The full-speed short-circuit time constants of synchronous machines are valid down to speeds of only a few per cent. At very low speeds the armature time constant separates into two different quantities applying separately to the direct and quadrature-axis and depending also on the rotor resistances.

Effect of Load or Tie Line

The results discussed previously apply to a synchronous machine connected to an infinite bus or short-circuited. In this section there are given the results of calculations with external armature inductance and resistance. The inductance may represent an extreme case of a reactive tie line and the resistance may represent a load.

Figures 9–12 show the results obtained. Only the machine with an amortisseur was considered. Figures 9 and 10, for an external reactance of unity, show that for this case the full-speed values are valid down to even smaller speeds with external reactance. Figures 11 and 12, for an external resistance of unity, show that for this case the full-speed values are only approximated at any speed. The field decrement P_1 changes rapidly, and the armature decrement is valid only above about 55 per cent speed.

The formulas of equations 5–8 may be used to compare the exact full-speed time constants with the conventional approximate values for the case of an external reactance (Figures 9 and 10), but for an external resistance (Figures 11 and 12) it is necessary to use the formulas

$$T_{ds}'=T_{do}'\left(\frac{x_d'x_q+r_a^2}{x_dx_q+r_a^2}\right) \qquad (14)$$

$$T_{ds}''=T_{do}''\left(\frac{x_d''x_q+r_a^2}{x_d'x_q+r_a^2}\right) \qquad (15)$$

$$T_{qs}''=T_{qo}''\left(\frac{x_q''x'_d+r_a^2}{x_qx_d'+r_a^2}\right) \qquad (16)$$

The first formula for the short-circuit field time constant T_{ds}' in terms of the field open-circuit time constant T_{do}', is well known, the second and third, for the direct T_{ds}'' and quadrature T''_{qs} axis subtransient time constants in terms of the corresponding open-circuit values T_{qo}'' and T_{do}'' are written simply by analogy to the first. All the exact values are found to check the approximate values within a few per cent at full speed.

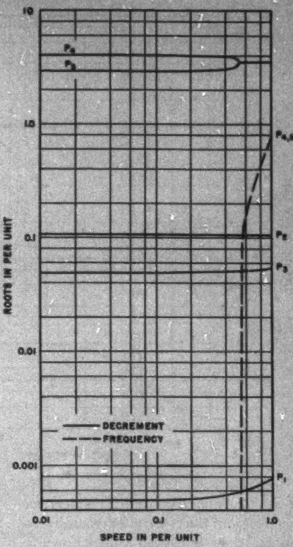

Figure 12. Synchronous machine—decrement factors and natural frequencies

(Roots of the characteristic equation 9)

With amortisseur

$x_l=0.2$ $x_f=0.1$
$x_{aq}=0.6$ $r_a=1.005$
$x_{ad}=1.0$ $r_{kq}=0.04$
$x_{kq}=0.2$ $r_{kd}=0.02$
$x_{kd}=0.1$ $r_f=0.00055$

For example, by the approximate formulas, with $r_a = 1.005$, $x_d' = 0.2909$, $x_d'' = 0.2476$ and $x_q'' = 0.35$,

$$T_{ds}''=9.28=\frac{1}{0.1078}$$

$$T_{qs}''=17.9=\frac{1}{0.0559}$$

$$T_{ds}'=631 \text{ or } 1262$$

$$=\frac{1}{0.001585} \text{ or } \frac{1}{0.000792}$$

All of these are in good agreement with Figures 11 and 12.

For the case of Figures 9 and 10, we similarly find

$$T_{ds}'=587 \text{ or } 1174$$

$$=\frac{1}{0.00170} \text{ or } \frac{1}{0.00085}$$

Design of Induction Heating Coils for Cylindrical Magnetic Loads

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Synopsis: Equations are presented whereby the authors' previous paper[1] is extended to the design of coils for heating magnetic loads, where the load remains in a fixed position relative to the coil during the heat. In heating magnetic materials, there is considerable variation in impedance of the coil circuit as the load temperature changes. The present paper first develops certain general principles by which a variable impedance circuit may be designed to absorb a predetermined maximum power, on the basis of constants determining the maximum and minimum impedance. Methods then are given by which these constants may be calculated for carbon steel loads of types SAE 1015 to SAE 1045, or their equivalent. In practice, these methods apply to most through heating and some surface heating applications. The accuracy of calculations is substantiated experimentally.

With the advent of considerable application of magnetic induction in heating steel for forging, it becomes more essential that the designer is able to precalculate the power, kilovolt-amperes, and efficiency variations during the heat in order to determine the required coil turns, capacitor values, and heating rates. It is the purpose of the present paper to outline methods by which these variations may be calculated and coil turns and capacitor values determined.

Numerous attempts have been made to adapt the theory of the induction heating of nonmagnetic loads to magnetic loads by assuming a constant permeability. However, unless some means is given of determining the value of permeability to use, the results are ambiguous. Baker[2] has shown that a value

$$\mu = 1.8B_M/H_M$$

where B_M is the maximum flux density (taken in Baker's paper as 18,000 lines per square centimeter) and H_M is the peak magnetic intensity at the outside of the load may be used to calculate the power input to thin moving steel strip. Kinn[3] proposes an almost identical formula for permeability for use with loads of various shapes. His value for B_M, however, is 16,000 lines per square centimeter. Kinn's paper is confined mainly to practical applications and, therefore, does not give theoretical derivation or experimental confirmation of the formulas supplied.

The conventional induction heating circuit used with rotating generators is shown in Figure 1. The purpose of the capacitor is to adjust the power factor at the generator terminals. In heating operations where the load remains in fixed position during the heat, the capacitor usually is adjusted to give a power factor at generator terminals in the range of unity to 0.9 leading while the load is in the magnetic state. As the load temperature rises above the Curie or critical point, there is an increase in current and lagging kilovolt-amperes to the induction heating coil.

If the generator voltage output is held constant by means of a regulator, usually the power output will decrease, but the increase in lagging kilovolt-amperes to the induction heating coil may cause an increase in generator output current to a value above its rating. In this case it is necessary to add capacitance in order to keep the generator output current within its rating. In typical applications the power output after the load temperature passes critical may decrease to as low as 30 per cent of the peak output occurring during the heat.

If there is no voltage regulator on the generator, as the load temperature rises above the critical point the resulting lagging power factor load on the generator causes a decrease in the generator output voltage. In this case it is desirable to add capacitance to keep the generator output voltage up to a value near its rating.

Usually, the number of coil turns is so selected that with rated generator terminal voltage the maximum power absorption during the heat is equal to the generator output rating. However, in cases where the power at the end of the heat is considerably less than the peak during the heat for a given coil voltage, fewer coil turns may be used, resulting in a maximum power absorption equal to generator output rating with a voltage less than the generator rating. In this case the generator output voltage is reduced by operating with a lagging power factor while the load is below the critical point. Then the generator output voltage is increased by addition of capacitance after passing the critical point. By this means it is sometimes possible to realize a higher average power during the heat and hence reduce heating time compared to that which would be obtained with a coil designed to absorb a maximum power output equal to generator rating at rated voltage.

Changes as Load Heats

In the definitions which follow, A, B, C, D, E, F, and G are defined implicitly to save space. For example, $R_P = N^2A$ is equivalent to $A = R_P/N^2$.

N = number of turns
$R_P = N^2A$ = resistance of inductor coil itself (ohms)
$R_S = N^2B$ = resistance of load as reflected in inductor coil (ohms)
$R = N^2(A+B) = N^2F$ = resistance of inductor coil with load (ohms)
$X_P = N^2C$ = reactance caused by flux between inside and outside radius of inductor coil (ohms)
$X_S = N^2D$ = reactance caused by flux within the load as reflected in inductor coil (ohms)
$X_a = N^2E$ = reactance caused by flux in air gap between coil and load (ohms)
$X = N^2(C+D+E) = N^2G$ = reactance of inductor coil with load (ohms)
E_L = voltage across inductor coil terminals AA (volts)
I_L = current through inductor coil (amperes)
P_L = power input to inductor coil and load (watts)
P_S = power dissipated in load (watts)

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These may be compared to P_1 of Figures 9 and 10.

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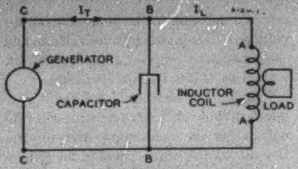

Figure 1. Conventional induction heating circuit using motor generator set

The required number of turns for any given total power to the coil and for any given terminal voltage may be calculated by means of equation 2.

In analyzing the changes which occur in a variable impedance circuit, it is convenient to represent the variation of reactance with resistance by a plot in rectangular co-ordinates. Referring to Figure 2, the total resistance R and reactance X of an induction heating coil with its load at room temperature are represented by the co-ordinates of point P_1. Since R_p, X_p, and X_a are nearly constant, the principal changes which occur in R and X as the load heats are caused by variations in the load resistance and reactance, R_S and X_S. When the power first is applied, R_S and X_S increase because of the increasing resistivity of the load with rising temperature. The load impedance, therefore, follows a locus such as $\overline{P_1P_2}$.

After a portion of the load reaches the critical temperature, R and X drop off rapidly. Let P_3 represent the impedance calculated by taking the entire load to be above the critical temperature and nonmagnetic with resistivity corresponding to final conditions. The assumption will be made that after the point of maximum impedance is reached R and X follow the straight line locus P_2P_3. This assumption has been checked experimentally and found to be reasonably accurate.

Suppose E_L, the inductor coil terminal voltage, is held constant as the load heats. The locus of inductor coil current I_L corresponding to impedance locus P_2P_3 is the circular arc I_2I_3.[4] The center of the circle OI_2I_3 lies on the line OQ, which makes the same angle α with the vertical as P_2P_3 makes with the horizontal. The power factor angle θ is equal to the angle OP makes with E_L. With conditions as shown in Figure 2, maximum power to the coil occurs at the point of maximum impedance, represented by P_2, since the component of inductor current that is in phase with E_L is greatest at this point.

The current locus of Figure 2 would result from a relatively large air gap reactance X_a. Under this condition the current through the inductor coil does not change greatly during the heating operation. However, if the air gap reactance is small, the magnitude of this current may change sufficiently so that maximum power to the coil occurs when the impedance is less than the maximum value. With the impedance locus shown in Figure 3, maximum power occurs when I_L is represented by OI_4, where Q is the center of the current locus and QI_4 is parallel to E_L. By plane geometry $\theta_4 = \angle QI_4O$, and $90°+\alpha = \angle QI_4O + \angle QOI_4 = 2\angle QI_4O$. Therefore,

$$\theta_4 = 45° + \alpha/2 \qquad (1)$$

The angle which OP_4 makes with E_L may be determined, therefore, by plotting P_2P_3 and measuring the angle α which this line makes with E_L. The intersection P_4 of OP_4 with P_2P_3 gives the values of R and X for maximum power to the coil. If θ_4 as defined by equation 1 is less than θ_2 or greater than θ_3, the intersection of OP_4 with P_2P_3 extended does not represent an impedance which actually occurs during the heating operation. If θ_4, or $45° + \alpha/2$, is less than θ_2, as in Figure 2, maximum power corresponds to maximum impedance. If θ_4 should be greater than θ_3, maximum power would occur at the end of the heating operation where the impedance is minimum.

Since F and G are proportional to the total resistance and reactance of the coil ($R = N^2F$ and $X = N^2G$), P_2P_3 in Figures 2 and 3 represents not only the locus of X versus R but, to a different scale, the locus of F versus G. The values of F and G used to determine the number of turns usually are based on conditions for maximum power to the coil. For $\theta_2 > 45° + \alpha/2$ (Figure 2), maximum power to the load corresponds to maximum impedance, or the values of F and G plotted at P_2. For $\theta_2 < 45° + \alpha/2$ (Figure 3), maximum power to the load occurs when F and G are determined by the co-ordinates of P_4.

Circuit Equations and Their Limitations

For any given total power P_L to the coil and coil terminal voltage E_L, the required number of turns is

$$N = E_L\sqrt{\frac{F}{P_L(F^2+G^2)}} \qquad (2)$$

By definition,

$$F = A + B \qquad (3)$$

$$G = C + D + E \qquad (4)$$

Formulas are given in the authors' previous paper[1] for the calculation of A, C, and E, and for the calculation of the load resistance and reactance coefficients B and D at the end of the heating operation, if the load is in the nonmagnetic state. The same limitations as given previously apply in the use of these formulas. These equations for resistance and reactance coefficients for nonmagnetic loads may be applied with reasonable accuracy to SAE 1010 to 1045 steel above 1,600 degrees Fahrenheit. For convenience, curves of resistivity versus temperature for SAE 1010 and SAE 1045 steels are given in Figure 4.

For the load in the magnetic state and at the temperature for maximum impedance, B and D may be calculated by

$$B = \left(\frac{3.14a_oK_1^2}{L}\right)\sqrt{f}K_7 \times 10^{-8} \qquad (5)$$

$$D = 0.65B \qquad (6)$$

K_7 is plotted against $P_S/(3.14\,a_oL)\sqrt{f}$, the watts per square inch of cylindrical surface of load within coil divided by the square root of frequency, in Figure 5. The ranges of extrapolation are indicated by dotted lines. For convenience the function

$$K_8 = \sqrt{f}\,K_7 \qquad (7)$$

is given as a function of $P_S/3.14\,a_oL$ for f = 10,000, 3,000, and 1,000 cycles per second in Figure 6. To determine B and D, it is first necessary to find P_S. The portion of the total coil power input absorbed by the load is $B/F = B/(A+B)$. Therefore, the power to the load is

$$P_S = \left(\frac{B}{A+B}\right)P_L \qquad (8)$$

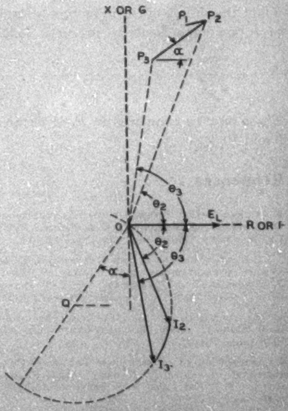

Figure 2. Impedance and current changes as load heats

B and F cannot be determined until P_g is known. However, the ratio B/F usually lies within the range 0.80 to 0.95. P_g, therefore, may be calculated first from equation 8 on the basis of any reasonable value of this ratio. An approximate value for $\sqrt{f}\,K_7$ or K_8 then can be obtained, which may be substituted into equation 5 to obtain a first approximation for B. Actually, a three per cent error in estimating P_g results in only a one per cent error in B, so that this first approximation is often sufficiently accurate for practical purposes. However, if greater accuracy is desired, the first approximation for B may be used to obtain a more accurate value of $B/(A + B)$ and the foregoing process repeated.

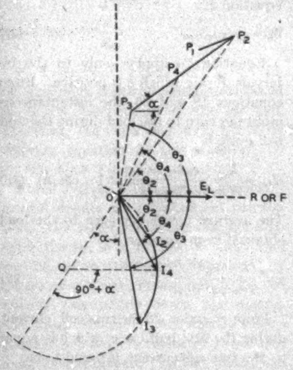

Figure 3. Impedance and current changes as load heats

LIMITATIONS

1. Taking the resistivity of the load as 40×10^{-6} ohm-inch at the temperature for maximum impedance, the maximum depth which magnetic flux and current penetrate into the load, or the depth of zone in which energy is induced (from equation 34) is

$$\delta_M=\frac{0.064}{K_6}\text{ (inches)} \tag{9}$$

If the load is hollow, its wall thickness should be at least as great as δ_M for the given equations to apply.

2. For an accuracy of plus or minus ten per cent, the ratio a_o/δ_M should be at least 8.0.

3. The correction for coil shortness should depend on the ratio of inside diameter of the induction heating coil to its length, the ratio of the outside diameter of the load to the inside diameter of the coil, and probably the ratio

$$\sqrt{R_g{}^2+X_g{}^2}/X_o=\sqrt{B^2+D^2}/E$$

of impedance associated with flux in the load to impedance associated with flux in the air gap. If this ratio is of the order of magnitude of 1.0 or greater, the flux through the load is appreciable in relation to the flux in the air gap, and the total flux is therefore considerably greater than the flux through the air gap alone. Consequently, a larger number of ampere turns is required to overcome the flux reluctance drop in the portion of the magnetic circuit external to the coil, or for any given number of ampere turns the magnetic intensity inside the coil is less than would result with a nonmagnetic load of the same dimensions. However, in the majority of applications in which a steel load is heated with a multiturn coil the effect of coil shortness is so small that it is believed that the equations given may be used with reasonable accuracy. The constant K_1, by which the actual ampere turns per inch must be multiplied to obtain the ampere turns per inch which would produce the same average power density if the coil were infinitely long, is derived (for nonmagnetic loads) in the authors' previous paper.[1]

Derivation of Expressions for B and D

The equations for B and D given in the foregoing text are based on experimental data and relations derived and checked experimentally under "More Exact Derivations," by means of which these data may be extended to other frequencies. The "Approximate Theory" is included to provide an understanding of certain physical principles in the induction heating of magnetic materials and to establish limitations for small diameters and wall thicknesses. The assumption made in the latter section does not limit the generality of the remainder of this paper.

Let

Subscript o refer to outside of load
Subscript M denote maximum values
δ = distance from outside of load (centimeters)
a_o = outside diameter of load (centimeters)
r = distance from axis of load (centimeters)
i = instantaneous current density in load (abamperes per square centimeter)
I_{ins} = instantaneous current in coil (abamperes)
I = rms current in coil (abamperes)
ϕ = instantaneous magnetic flux within radius r (maxwells)
e = instantaneous voltage per turn to bring about time rate-of-change of flux within load $= d\phi_o/dt$ (abvolts)
β = magnetic flux density (maxwells per square centimeter)
H = magnetic intensity
ρ = resistivity (abohms per centimeter)
t = time (seconds)
f = frequency (cycles per second)
$\omega=2\pi f$
L = length of coil (centimeters)
N = number of turns

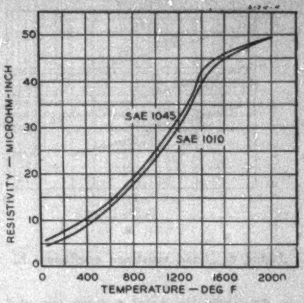

Figure 4. Resistivity versus temperature for SAE 1010 and SAE 1045 steel

The total flux within a circle of radius r is, by definition,

$$\phi=2\pi\int_0^r r\beta dr \tag{10}$$

By Ohm's law, with positive directions as indicated in Figure 7,

$$2\pi r\rho i=\frac{\partial\phi}{\partial t} \tag{11}$$

Providing, as will be assumed, that the radius of the load is sufficiently great so that r is nearly equal to the outside radius over the region in which β and i are appreciable, equation 10 may be written

$$\begin{aligned}\phi&=\pi a_o\int_0^r\beta dr\\&=\pi a_o\int_\delta^{a_o/2}\beta d\delta\end{aligned} \tag{12}$$

where $\delta=a_o/2-r$ is the distance from the outside of the load. Equation 11 may be written

$$\rho\pi a_o i=\partial\phi/\partial t \tag{13}$$

Consider only coils and loads of large length to diameter ratios. Then, as in the induction heating of nonmagnetic materials, the magnetic intensity is equal

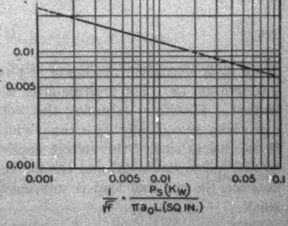

Figure 5. Load resistance coefficient K_7

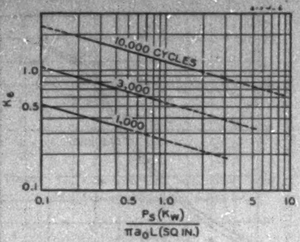

Figure 6. Load resistance coefficient K_s

to 4π times the total current per unit axial length outside the given radius. The current per unit axial length in the coil is $(N/L)\,I_{ins}$. Therefore,

$$H=4\pi\left[\frac{N}{L}I_{ins}-\int_0^{\delta} i\,d\delta\right] \tag{14}$$

The minus sign before the integral results from the choice of positive direction for i (Figure 7). Differentiating equation 13 with respect to δ and substituting equation 12 for ϕ,

$$\frac{\partial i}{\partial \delta}=-\frac{1}{\rho}\cdot\frac{\partial \beta}{\partial t} \tag{15}$$

Approximate Theory

Most induction heating applications require inductor coil current densities such as to produce high saturations near the outside of the load. Therefore, in the present section assume the $\beta-H$ curve shown in Figure 8, in which $\beta=\pm\beta_M$ or zero. Positive directions are indicated in Figure 7. Take the instant $t=0$ when the current in the inductor coil is zero after having been negative. Assume that at this instant the flux density β is negative (or to the left) between $\delta=0$ and $\delta=\delta_M$, where δ_M is the maximum depth of penetration to be determined later. The flux through the load at $t=0$ is

$$\phi_o|_{t=0}=-\pi a_o\delta_M\beta_M \tag{16}$$

The portion of the voltage per turn utilized in bringing about a time rate-of-change of flux within the load is

$$e=\frac{d\phi_o}{dt} \tag{17}$$

Since ϕ_o is negatively maximum at $t=0$, e is zero at $t=0$, but positive after this instant. Therefore, assuming this voltage varies sinusoidally with time, which is the case at least if the voltage applied to the coil is sinusoidal and the air gap between the coil and load is sufficiently small,

$$e=e_M\sin\omega t \tag{18}$$

Equation 15 shows that at any instant the current density in the load varies with δ only where the flux density β is changing with time. By assumption, β can have only three values: β_M, $-\beta_M$, and zero. At any given time, i can change only where β is going from one to another of these values.

The magnetic intensity at the outside of the load is

$$H_o=4\pi(N/L)I_{ins} \tag{19}$$

Consequently, at the instant $t=0$ at which the current in the coil becomes positive, the direction of magnetization at the outside of the load also becomes positive. Denote the depth of positive magnetization by δ_+. The total flux within the load is

$$\phi_o=\pi a_o[\delta_+\beta_M+(\delta_M-\delta_+)(-\beta_M)]=\pi a_o\beta_M(2\delta_+-\delta_M) \tag{20}$$

Integrating equation 17 and substituting equation 18 for e,

$$\begin{aligned}\phi_o&=\int_0^t e\,dt+\text{constant}\\&=-(e_M/\omega)\cos\omega t+\text{constant}\end{aligned} \tag{21}$$

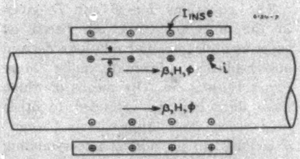

Figure 7. Current and flux directions—positive

Combining equations 20 and 21, and noting that $\delta_+=0$ at $t=0$,

$$\delta_+=(e_M/2\pi a_o\beta_M\omega)(1-\cos\omega t) \tag{22}$$

δ_+ is plotted against time in Figure 9.

The time rate-of-change of flux $\partial\phi/\partial t$ is independent of position for $\delta<\delta_+$, and zero for $\delta>\delta_+$. Therefore, from equation 13,

$$\left.\begin{aligned}i&=\frac{1}{\rho\pi a_o}\cdot\frac{d\phi_o}{dt}&&\text{for }\delta<\delta_+\\&=0&&\text{for }\delta>\delta_+\end{aligned}\right\} \tag{23}$$

The total current per unit length in the load is

$$\int_o^{a_o/2} i\,d\delta=i\delta_+ \tag{24}$$

β is zero for $\delta>\delta_M$ by assumption. Therefore, H is also zero. Substituting $\delta=a_o/2$ into equation 14, therefore, gives

$$I_{ins}=(L/N)i\delta_+ \tag{25}$$

Substituting equations 23 and 21 for i and equation 22 for δ_+,

$$I_{ins}=\frac{L}{N}\cdot\frac{e_M^2}{2\pi^2\rho a_o^2\beta_M\omega}\sin\omega t(1-\cos\omega t) \tag{26}$$

$e=0$ at $\omega t=\pi$. Therefore, from equations 17 and 13, $d\phi/dt=0$ and the current density i is zero; i was also zero at $t=0$. The conditions at $\omega t=\pi$, therefore, are identical (except for directions) with the conditions assumed at $t=0$. During the remaining half cycle a wave of negative magnetization enters the load. The phenomena of the first half cycle are repeated with changes in polarity, as shown in Figure 9. The maximum depth of penetration of electrical energy is obtained by substituting $\omega t=\pi$ into equation 22.

$$\delta_M=e_M/\pi a_o\beta_M\omega \tag{27}$$

Equation 26 applies only to the intervals during which e is positive. From equations 18 and 26, the instantaneous input per turn to the load during the first half cycle is

$$eI_{ins}=\frac{L}{N}\cdot\frac{e_M^3}{2\pi^2\rho a_o^2\beta_M\omega}\sin^2\omega t(1-\cos\omega t) \tag{28}$$

The average power per turn to the load obtained from equation 28 is

$$w=\frac{L}{N}\cdot\frac{e_M^3}{4\pi^2\rho a_o^2\beta_M\omega} \tag{29}$$

From equation 26 the rms coil current during the first half cycle, which is equal to the rms coil current, is found to be

$$I=\sqrt{\frac{5}{8}}\cdot\frac{L}{N}\cdot\frac{e_M^2}{2\pi^2\rho a_o^2\beta_M\omega} \tag{30}$$

Comparing equations 29, 18, and 30, the ratio of abwatts to abvolt-amperes or watts to volt-amperes is

$$\frac{w}{0.707e_MI}=\sqrt{0.8}=0.895 \tag{31}$$

Solving equation 30 for e_M and substituting into equation 29, the power to the load per turn is

$$w=3.16a_o\sqrt{\rho\beta_M\omega}\cdot\sqrt{\frac{N}{L}}\cdot I^{1.5} \tag{32}$$

or the total power to the load is

$$Nw=7.92a_o\sqrt{\frac{\rho\beta_Mf}{L}}\cdot(NI)^{1.5} \tag{33}$$

Table I

	Kilowatts Per Square Inch to Load	Ratio (Watts Divided By Volt-Amperes) to Load
Calculated	0.244	0.895
Measured	0.239	0.892

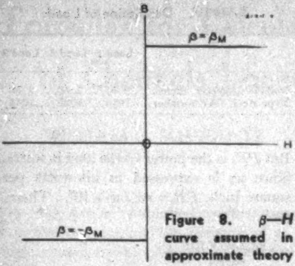

Figure 8. β—H curve assumed in approximate theory

Making use of equations 27, 30, 33, 5, and 7 and converting all quantities to the ohm-inch system, it can be shown that

$$\delta_M = 1.595\times 10^3 \frac{\rho}{K_6} \tag{34}$$

For a given density of power to the load, the maximum depth of penetration should be nearly independent of the air gap. Therefore, the expression used for K_6 in deriving equation 34 was obtained by putting $K_1 = 1$ in equation 5.

Table I gives a comparison of experimental results with theoretical calculations, the latter based on the approximate theory. An SAE 1015 steel load was heated to 210 degrees Fahrenheit by means of a closely fitting coil. The frequency was 10,000 cycles per second. The measured coil current was such as to produce 486 ampere turns (rms) per inch axial length of coil. In the theoretical calculation, the resistivity of SAE 1015 steel at 210 degrees Fahrenheit and the flux density β_M were taken respectively as 6.75×10^{-6} ohm-inches and 18,500 lines per square centimeter.

The following conclusions may be drawn from the analysis so far presented:

1. The total depth of penetration of electrical energy in the magnetic state is usually small. Table II gives values of δ_M corresponding to two kilowatts per square inch to the load and the temperature for maximum impedance with constant current; ρ is taken as 40×10^{-6} ohm-inch, and K_6 in equation 34 is obtained from Figure 6.

2. The power density is approximately proportional to the 1.5 power of current. This relation was observed by Rosenberg.[5,6,7] As noted by Baker,[2] however, Rosenberg's explanation is not very satisfactory.

3. It frequently has been assumed that the resistance and reactance of an induction heating coil with a magnetic load could be calculated by assuming "some" constant value of permeability. This is not strictly true,since any constant value of permeability would result in a ratio of watts to volt-amperes to the load of 0.707 at sufficiently high frequencies.

4. With power densities as used in induction heating practice and with δ_M small compared to the load diameter, the load power caused by hysteresis is usually negligible. For instance, considering a 1-inch diameter load of SAE 1020 steel with a power density of 500 watts per square inch of load surface at 10,000 cycles, the power caused by hysteresis would be less than 22 watts per square inch of load surface, or four per cent of the total power. This is based on an assumed maximum flux density of 100,000 lines per square inch and a hysteresis energy of 0.05 joule per cubic inch per cycle,[10] which is from data with the steel at room temperature. Further data indicate a decrease in hysteresis loss as temperature is increased, the hysteresis loss approaching zero as the critical temperature is approached.[9]

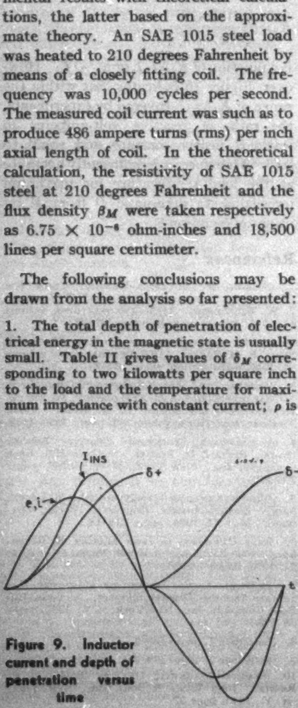

Figure 9. Inductor current and depth of penetration versus time

More Exact Derivations

The saturation flux densities of carbon steels usually treated by induction heating are known at present only at room temperature. In order to design coils for heating above the critical point, however, the values of the load resistance and reactance coefficients B and D must be determined at the temperature of maximum impedance, about 1,200 degrees Fahrenheit. The present section provides a means of accomplishing this, and at the same time affords a more accurate theory which avoids the assumption that $\beta = \pm$ constant. Use is made of experimental information which is readily obtainable.

Again consider loads of sufficient radius so that the distance from the axis may be taken as the outside radius $a_0/2$ at all points for which the magnetic intensity H, the flux density β, and the current density i are appreciable. Neglect hysteresis. Then for any given value of H there exists one and only one value of β or, mathematically,

$$\beta = \phi_{(H)} \tag{35}$$

Differentiating equation 14 twice with respect to δ and substituting into equation 15,

$$\frac{\partial^2 H}{\partial \delta^2} = \frac{4\pi}{\rho}\cdot\frac{\partial \beta}{\partial t} \tag{36}$$

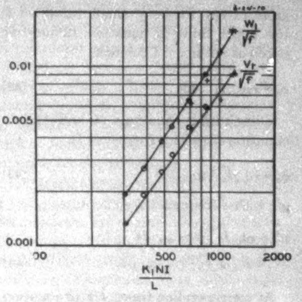

Figure 10. Plot of experimental data

⊙—Coil 1, 10,000 cycles
+—Coil 2, 10,000 cycles
×—Coil 3, 3,000 cycles

Making use of equation 35,

$$\frac{\partial^2 H}{\partial \delta^2} = \frac{4\pi}{\rho}\mu'_{(H)}\frac{\partial H}{\partial t} \tag{37}$$

where $\mu'_{(H)}$ is the differential permeability defined by

$$\mu' = \frac{dB}{dH} \tag{38}$$

Since for any given distance δ from the outside of the load and time, there exists one and only one value of H:

$$H = \psi_{(\delta,\, t)} \tag{39}$$

As $\psi_{(\delta,\, t)}$ is a solution of equation 37, it can be shown that $H' = \psi_{(k\delta,\, k^2t)}$ is also a solution. At the outside of the load $\psi_{(\delta,\, t)} = \psi_{(o,\, t)}$ and $\psi_{(k\delta,\, k^2t)} = \psi_{(o,\, k^2t)}$. Suppose for a given peak magnetic intensity H_{OM} at the outside of the load and a given frequency f_1, H is represented by $\psi_{(\delta,\, t)}$. Then for the same magnetic intensity H_{oM} at the outside of the load and a frequency $k^2 f_1$, the magnetic intensity is represented by $\psi_{(k\delta,\, k^2t)}$.

With a frequency f_1,

$$\beta = \phi_{(H)} = \phi(\psi_{(\delta,\, t)}) = \Omega_{(\delta,\, t)} \tag{40}$$

With the same peak magnetic intensity at the outside of the load and a frequency $k^2 f_1$,

$$\beta = \phi_{(H')} = \phi(\psi_{(k\delta,\, k^2t)}) = \Omega_{(k\delta,\, k^2t)} \tag{41}$$

Since the flux density is assumed negli-

Table II

Frequency in Cycles Per Second	Depth of Penetration in Inches
10,000	0.066
3,000	0.145

gible except for very small values of δ, the upper limit in equation 12 may be taken as infinity. Therefore,

$$\phi_0=2\pi a_0\int_0^\infty \mathfrak{B}d\delta \tag{42}$$

Substituting equations 40 and 41 into equation 42, with a frequency f_1,

$$\phi_0=\pi a_0\int_0^\infty \Omega_{(\delta,\,t)}d\delta \tag{43}$$

or, with a frequency k^2f_1,

$$\begin{aligned}\phi'_0&=\pi a_0\int_0^\infty \Omega_{(k\delta,\,k^2t')}d\delta\\&=\pi a_0/k\int_0^\infty \Omega_{(\delta,\,k^2t')}d\delta\end{aligned} \tag{44}$$

At corresponding times, k^2t' in equation 44 is equal to t in equation 43. Therefore, the total flux in the load is proportional to $1/k$ or $1/\sqrt{f}$. The load impedance is proportional to the product of the magnitude of this flux multiplied by frequency, or

$$(1/\sqrt{f})f=\sqrt{f}$$

Let

w_1 = kilowatts per square inch to the load

v_1 = reactive kilovolt-amperes per square inch to the load

K_1NI/L = effective ampere turns per inch (rms)

Then from the foregoing reasoning, w_1 and v_1 may be written in the form

$$w_1=\sqrt{f}\,F_1(K_1NI/L) \tag{45}$$

$$v_1=\sqrt{f}\,F_2(K_1NI/L) \tag{46}$$

where the functions F_1 and F_2 are independent of frequency. To determine these functions for SAE 1015 steel at the load temperature for maximum impedance and to show experimentally that $w_1/\sqrt{f}$ and $v_1/\sqrt{f}$ are independent of frequency for any given K_1NI/L, data were taken at 10,000, 9,600, and 3,000 cycles per second. The results are shown in Figure 10. The variations in $w_1/\sqrt{f}$ and $v_1/\sqrt{f}$ between 9,600 and 10,000 cycles per second were negligible for a given K_1NI/L. Dimensions of coils and loads are given in Tables III and IV. In each reading the current through the induction heating coil was held constant by varying the voltage. This current was measured through a current transformer by a dynamometer type ammeter. Readings were taken at the instant of maximum coil terminal voltage. The power to the coil also was found to be maximum at this instant. To measure this power as accurately as possible, the capacitor at BB (Figure 1) was adjusted for approximately unity power factor at the generator terminals CC for the instant readings were taken. The generator output was measured by a wattmeter connected at CC. The power to the load was obtained by subtracting the calculated coil copper loss.[1] After making a slight correction for conductor wall thinness at 3,000 cycles per second (no correction was required at 10,000 cycles per second), the maximum ratio of calculated coil I^2R loss to generator output was 0.04. Capacitor and bus losses between the wattmeter and coil terminals were negligible. The voltage at the terminals of the induction heating coil was measured by a voltmeter connected at AA. It was assumed that the reactive volt-amperes to the coil could be determined from the watts and volt-amperes by the usual method. Experience indicates that, after making allowance for the reactance of the bus, the reactive volt-amperes so calculated agree closely with the volt-amperes of capacitance required to balance to unity power factor at the generator terminals, even though the wave form of the current through the inductor coil is not truly sinusoidal. The reactances associated with the time rate-of-change of flux in the coil copper and air gap are, respectively, N^2C and N^2E. The reactive volt-amperes to the load were obtained by subtracting $I^2N^2(C+E)$ from the reactive volt-amperes to the coil. The watts per square inch and vars per square inch to the load then were found by dividing the power and reactive volt-amperes by the load surface within the coil πa_0L.

With a frequency of 9,600 cycles only, the foregoing experiments were repeated with SAE 1020 and SAE 1045 loads. No significant differences could be detected from the results with SAE 1015 steel. At 1200 degrees Fahrenheit, the difference in resistivities between SAE 1015 and SAE 1045 steels is about three per cent. According to equation 33, if the magnetic properties of the two steels were identical, the load resistances should differ by about two per cent. This variation would be negligible in most induction heating calculations.

By definition,

$$B=R_s/N^2=I^2R_s/(NI)^2 \tag{47}$$

Table III. Description of Induction Heating Coils

	Coil 1	Coil 2
Total turns, N	35	19
Inside diameter, inches	3.29	3.20
Length over turns, L, inches	14.62	7.05
Conductor:		
Dimension parallel to axis of coil, inches	0.375	0.297
Dimension perpendicular to axis of coil, inches	0.25	0.25
Wall thickness, inches	0.045	0.048
Resistivity, microhm inches	0.73	0.73

Table IV. Description of Loads

	Load 1	Load 2	Load 3
Outside diameter, inches	2.51	2.50	2.50
Type steel, SAE number	1015	1020	1045

But I^2R_s is the power to the load in watts. Since w_1 is expressed in kilowatts per square inch, $I^2R_s=\pi a_0Lw_1\times 10^3$. Therefore,

$$\begin{aligned}B&=\frac{\pi a_0Lw_1\times 10^3}{(NI)^2}\\&=\frac{\pi a_0k_1^2}{L}\sqrt{f}\,K_7\times 10^{-3}\end{aligned} \tag{48}$$

where

$$K_7=\frac{w_1\times 10^6/\sqrt{f}}{(K_1NI/L)^2} \tag{49}$$

From equation 45 $w_1/\sqrt{f}$ is a function of $K_1(NI/L$, or K_1NI/L is a function of $w_1/\sqrt{f}$. Therefore, K_7 is also a function of $w_1/\sqrt{f}$. Similarly, it may be shown that

$$D=\frac{\pi a_0K_1^2}{L}\sqrt{f}\,K_8\times 10^{-3}$$

where

$$K_8=\frac{v_1\times 10^6/\sqrt{f}}{(K_1NI/L)^2}$$

is a function of K_1NI/L or $w_1/\sqrt{f}$. However, Figure 10 indicates that within the range of K_1NI/L covered by experiments, the ratio of reactive volt-amperes to watts is nearly constant and equal to 0.65. Therefore, it is sufficiently accurate at the load temperature for maximum impedance to determine D from B by means of equation 6.

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Formulas for Calculating the Inductance of Channels Located Back to Back

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In 1937 Arnold[1] published an analytic procedure enabling the designer of a single-phase 2-conductor bus to determine combinations of conductor spacing and cross section that render proximity and skin effects (a-c resistance, eddy-current losses, and current density distortion) a minimum. For explicit details the reader is referred to Arnold's paper. For our purpose it suffices to remark:

1. In general the designer is afforded an unlimited number of combinations of conductor spacing and cross section.
2. As the theoretical cross sections are infinitely thin, it is necessary, in practice, to use finitely thick structural shapes whose mean perimeters approximate the theoretical shapes.
3. As the ratio of conductor spacing to perimeter of theoretical cross section increases, the approximately equivalent structural shape associated with a particular spacing shifts from rectangular strap to channel to circular or square tubular conductor.
4. Although the infinitely thin theoretical cross sections produce zero proximity and skin effects, such is not true of the finitely thick conductors actually used: for though the proximity effect is effectively yet zero, unavoidably there is introduced minimum skin effect.
5. The skin effect losses, and hence the a-c resistance, can be calculated from a given simple formula.
6. The accuracy of this formula and the correctness of the general theory are verified by experimental data set forth in Arnold's paper.

Despite, however, the availability and the obvious usefulness of this general theory, it would seem that no considerable use, if any, has yet been made of it. In some part, no doubt, this neglect is due to lack of knowledge of it: Arnold's paper was published in the *Journal* of the Institution of Electrical Engineers, a British periodical which in this country enjoys but little circulation outside of engineering libraries or the technical departments of a few of the larger public libraries. In other part, however, this neglect doubtlessly can be attributed to the fact that, largely, a-c industrial-bus design is polyphase and has as its desideratum the achievement of minimum inductance and hence minimum reactance voltage drop. Designs achieving this end involve closely spaced conductors and thus greater than minimum eddy-current losses, whereas practical designs achieving minimum eddy-current losses involve fairly widely spaced conductors and thus increased reactance voltage drop.

When, however, single-phase busses are to be designed and minimum reactance is not so much the desideratum as is minimum eddy-current loss (for example, very heavy current busses where eddy-current loss is the predominant factor in design), Arnold's theory affords preferred designs. In consequence, it is of interest to the designer of such busses that there be available to him formulas enabling calculation of the electrical and mechanical performance of designs stemming from Arnold's theory. This, as already mentioned, yields conductor cross sections to be approximated by standard structural shapes: rectangular strap, channel, and circular or square tubular conductors.

Now formulas for calculating the a-c resistance, a-c inductance, mutual electromagnetic forces on short circuit, and other conductor parameters essential to knowledge of the electrical and mechanical (short-circuit) performance of the bus are available for rectangular strap and for circular and square tubular conductors.[2] As regards (uniformly thick) channel conductors located back to back, Arnold[1] has advanced a formula for calculating the a-c resistance and Higgins[3,4] has advanced formulas for calculating the mutual electromagnetic forces exerted between the conductors on short circuit. But explicit formulas for the inductance as a function of the geometrical parameters specifying conductor spacing and cross section have not been published hitherto. Yet, inasmuch as among the various combinations of conductor spacing and cross section afforded by Arnold's theory, that combination which affords the best all-around design is usually one comprised of moderately spaced channels located back to back, it is most desirable to have formulas for the inductance.

Such formulas are derived in this paper, it being assumed that the conductors are nonmagnetic, are of such length that end effects are negligible, are right-cornered, are uniformly thick, and carry currents uniformly distributed over their cross sections.

Of these five conditions, the first and second commonly are satisfied in practice. Right-corneredness, however, though essential to tractable analysis, is not always true in practice: channels built up of strap conductors are usually right-angled, whereas channels drawn, rolled, or otherwise formed in one piece usually have rounded edges. However, error introduced by neglecting the rounded edges is usually quite small; moreover, where necessary the increase in inductance can be approximated by use of semi-empirical formulas derived by Dwight and Wang.[5] Uniform thickness, on the other hand, is true both of channels built up of strap conductor and (except for possible slight variation in the immediate vicinity of a rounded edge) of most commercial channels. Finally, in designs based on Arnold's theory, proximity effect is practically obviated and skin effect minimized; though this minimum skin effect tends to produce an inductance slightly less than the d-c value, the actual change is practically negligible (see, for example, the discussion following reference 5 or the experimental results published in reference 6); whence the assumption of uniform current distribution is consistent with fact, introduces negligible error, and enables solution of an otherwise intractable problem.

If these five postulates are granted, derivation of the desired formulas, though lengthy, is not particularly difficult. The conductor cross sections can be considered, in various fashions, as combinations of parallel-sided rectangular areas. By virtue of the fundamental theorem of geometric mean distance theory, the inductance can be expressed in terms of the geometric mean distances between these rectangular areas. These geometric mean distances can be calculated from known formulas[7,8] as shown under "Some General Relationships." Accordingly, determination of the inductance as a function of the parameters

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defining the (spacing and cross sections of the conductors of the) bus reduces, in essence, to specific formulation of these geometric mean distances and substitution of them in the general equation for the inductance (see "Derivation of the General Formula"). To check the correctness of the resulting formula, we note that the channels can be (1) so proportioned or (2) so located relative to each other that the bus reduces to one comprised of rectangular strap conductors. Corroboratively, the known formula[9,10] for the inductance of this latter bus is yielded by the general formula for channels back to back ("Corroboration of Equation 10").

This general formula contains some 75 terms. A much simpler formula, similarly derived—but on the assumption that the conductors can be considered as comprised of line segments (Figure 5)—facilitates computation preliminary to final choice of cross section. To check the correctness of this formula, we note that if the flanges are taken to be of zero width, the bus reduces to one comprised of infinitely thin strap conductors. Corroboratively, the known formula[9,10] for the inductance of this bus is yielded by the general formula for line segment channels back to back "(Approximations to the General Formula").

The desired formulas having been obtained and checked, their use is illustrated by the numerical solution of a typical problem in bus design ("Some Illustrative Examples"). Additionally, this illustrative example reveals certain precautions that must be observed in the use of the approximate formulas based on line segment channels ("Remarks on the Approximate Formula").

Some General Relationships

From well-known theory we have that the inductance of a bus comprised of two long parallel cylindrical nonmagnetic conductors, each of uniform but otherwise arbitrary cross section, is

$$L=2\log\frac{D_{ab}^2}{D_{aa}D_{bb}} \tag{1}$$

wherein D_{aa} and D_{bb} are the self geometric mean distances of the conductor cross sections and D_{ab} is the geometric mean distance between the cross sections.

If each of the two conductors is comprised of several component conductors connected in parallel, m comprising that with cross section S_a and n that with cross section S_b, equation 1 can be written[8]

$$L=-\left(\frac{2}{I^2}\right)\sum_{i=1}^{m+n}\sum_{j=1}^{m+n}w_iw_jS_iS_j\log D_{ij} \tag{2}$$

wherein I is the conductor current; w_i the current density in the ith component of cross sectional area S_i, taken positive for current flow in one direction and negative for current flow in the opposite direction; and $D_{ij}\equiv D_{ji}$ is the geometric mean distance between areas S_i and S_j.

If $S_i=RS$ and $S_j=R'S'$ are two arbitrarily located parallel-sided rectangular areas (Figure 1), the geometric mean distance between them is given by

$$S_iS_j\log D_{ij}=-\left(\frac{25}{12}\right)S_iS_j-\left(\frac{1}{24}\right)\sum_{p=1}^{4}\sum_{q=1}^{4}(-1)^{p+q}K(A_p,B_q) \tag{3}$$

wherein

$$K(A_p,B_q)=(A_p^4-6A_p^2B_q^2+B_q^4)\times\log(A_p^2+B_q^2)^{1/2}-4A_pB_q^3\times\tan^{-1}\frac{A_p}{B_q}-4A_p^3B_q\tan^{-1}\frac{B_q}{A_p} \tag{4}$$

and

$$A_1=|D+R+R'|;\quad A_2=|D+R|;\quad A_3=|D|;\quad A_4=|D+R'| \tag{5}$$

$$B_1=|P+S+S'|;\quad B_2=|P+S|;\quad B_3=|P|;\quad B_4=|P+S'| \tag{6}$$

Derivation of the General Formula

Consider a bus comprised of two identical channels back to back (Figure 2). Each of the channels can be taken as comprised of two component conductors: one of full rectangular cross section wherein current exists in the same direction as in the channel; the other of smaller rectangular cross section having current in opposite direction. Hence $m=n=2$, and from equation 2,

$$L=-\left(\frac{2}{I^2}\right)\sum_{i=1}^{4}\sum_{j=1}^{4}w_iw_jS_iS_j\log D_{ij} \tag{7}$$

By definition of geometric mean distance, $D_{ij}\equiv D_{ji}$. By virtue of the geometric symmetries in Figure 2,

$$D_{11}=D_{44};\ D_{22}=D_{33};\ D_{12}=D_{34};\ D_{13}=D_{24}$$

Further,

$$S_1=S_4;\ S_2=S_3;\ w_1=-w_2=w_3=-w_4=I/(S_1-S_2)$$

In consequence of these relationships, we have from equation 7

$$L=-\left(\frac{4}{A^2}\right)[S_1^2(\log D_{11}-\log D_{14})+S_2^2(\log D_{22}-\log D_{23})-2S_1S_2(\log D_{12}-\log D_{13})] \tag{8}$$

wherein

$$A=(S_1-S_2)=[rs-(r-t)(s-2t)] \tag{9}$$

Corroboratively, if $S_2=S_3=0$, we have rectangular strap conductors of cross section $S_1=S_4$ and equation 8 reduces to the known formula,

$$L=4\log\frac{D_{14}}{D_{11}}$$

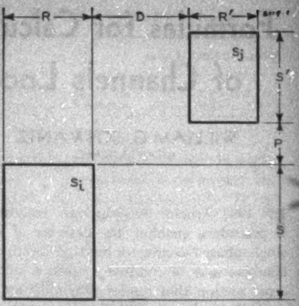


Figure 1. Relative location of two parallel-sided rectangles

Determining the six geometric mean distances in the right-hand member of equation 8 by use of equations 3–6—the explicit expressions are to be found in Appendix I—substituting appropriately in equation 8, and rearranging terms, we have

$$\begin{aligned}L=&-\left(\frac{1}{6}\right)[rs-(r-t)(s-2t)]^2\{4r^4\log r+\\&4s^4\log s-12t^4\log t-2d^4\log d+\\&4(r-t)^4\log(r-t)-4(s-t)^4\times\\&\log(s-t)+4(s-2t)^4\log(s-2t)+\\&4(d+r)^4\log(d+r)-4(d+2r)^4\times\\&\log(d+2r)-2(d+2t)^4\log(d+2t)+\\&4(d+r+t)^4\log(d+r+t)-2(r^4-\\&6r^2s^2+s^4)\log(r^2+s^2)-2(r^4-6r^2t^2+\\&t^4)\log(r^2+t^2)-2[(r-t)^4-6(r-\\&t)^2t^2+t^4]\log[(r-t)^2+t^2]-2[(s-t)^4-\\&6(s-t)^2t^2+t^4]\log[(s-t)^2+t^2]+\\&(d^4-6d^2s^2+s^4)\log(d^2+s^2)+2[(d+\\&t)^4-6(d+t)^2t^2+t^4]\log[(d+t)^2+t^2]+\\&2[r^4-6r^2(s-t)^2+(s-t)^4]\times\\&\log[r^2+(s-t)^2]+2[(r-t)^4-\\&6(r-t)^2(s-t)^2+(s-t)^4]\times\\&\log[(r-t)^2+(s-t)^2]-2[(r-t)^4-\\&6(r-t)^2(s-2t)^2+(s-2t)^4]\times\\&\log[(r-t)^2+(s-2t)^2]-2[(d+r)^4-\\&6(d+r)^2s^2+s^4]\log[(d+r)^2+s^2]+\\&[(d+2r)^4-6(d+2r)^2s^2+s^4]\times\\&\log[(d+2r)^2+s^2]-2[(d+r)^4-6(d+\\&r)^2t^2+t^4]\log[(d+r)^2+t^2]+2[(d+\\&2r)^4-6(d+2r)^2t^2+t^4]\times\\&\log[(d+2r)^2+t^2]-2[(d+r+t)^4-\\&6(d+r+t)^2t^2+t^4]\times\\&\log[(d+r+t)^2+t^2]+2[(d+r)^4-\\&6(d+r)^2(s-t)^2+(s-t)^4]\times\\&\log[(d+r)^2+(s-t)^2]-2[(d+2r)^4-\\&6(d+2r)^2(s-t)^2+(s-t)^4]\times\\&\log[(d+2r)^2+(s-t)^2]-2[(d+t)^4-\\&6(d+t)^2(s-t)^2+(s-t)^4]\times\\&\log[(d+t)^2+(s-t)^2]+[(d+2r)^4-\\&6(d+2r)^2(s-2t)^2+(s-2t)^4]\times\\&\log[(d+2r)^2+(s-2t)^2]+[(d+2t)^4-\end{aligned}$$

$$
\begin{aligned}
&6(d+2t)^2(s-2t)^2+(s-2t)^4]\times\\
&\log[(d+2t)^2+(s-2t)^2]+2[(d+r+\\
&t)^4-6(d+r+t)^2(s-t)^2+(s-t)^4]\times\\
&\log[(d+r+t)^2+(s-t)^2]-2[(d+r+\\
&t)^4-6(d+r+t)^2(s-2t)^2+(s-2t)^4]\times\\
&\log[(d+r+t)^2+(s-2t)^2]+16rs^3\times\\
&\tan^{-1} r/s+16r^3s\tan^{-1} s/r+16rt^3\times\\
&\tan^{-1} r/t+16r^3t\tan^{-1} t/r+\\
&16(r-t)t^3\tan^{-1}(r-t)/t+16(r-t)^3t\times\\
&\tan^{-1} t/(r-t)+16(s-t)t^3\times\\
&\tan^{-1}(s-t)/t+16(s-t)^3t\times\\
&\tan^{-1} t/(s-t)-8ds^3\tan^{-1} d/s-\\
&8d^3s\tan^{-1} s/d-16(d+t)t^3\times\\
&\tan^{-1}(d+t)/t-16(d+t)^3t\times\\
&\tan^{-1} t/(d+t)-16r(s-t)^3\tan^{-1} r\div\\
&(s-t)-16r^3(s-t)\tan^{-1}(s-t)/r-\\
&16(r-t)(s-t)^3\tan^{-1}(r-t)/(s-t)-\\
&16(r-t)^3(s-t)\tan^{-1}(s-t)/(r-t)+\\
&16(r-t)(s-2t)^3\tan^{-1}(r-t)\div\\
&(s-2t)+16(r-t)^3(s-2t)\times\\
&\tan^{-1}(s-2t)/r(-t)+16(d+r)s^3\times\\
&\tan^{-1}(d+r)/s+16(d+r)^3s\times\\
&\tan^{-1} s/(d+r)-8(d+2r)s^3\times\\
&\tan^{-1}(d+2r)/s-8(d+2r)^3s\times\\
&\tan^{-1} s/(d+2r)+16(d+r)t^3\times\\
&\tan^{-1}(d+r)/t+16(d+r)^3t\times\\
&\tan^{-1} t/(d+r)-16(d+2r)t^3\times\\
&\tan^{-1}(d+2r)/t-16(d+2r)^3t\times\\
&\tan^{-1} t/(d+2r)+16(d+r+t)t^3\times\\
&\tan^{-1}(d+r+t)/t+16(d+r+t)^3t\times\\
&\tan^{-1} t/(d+r+t)-16(d+r)(s-t)^3\times\\
&\tan^{-1}(d+r)/(s-t)-16(d+r)^3\times\\
&(s-t)\tan^{-1}(s-t)/(d+r)+16(d+\\
&2r)(s-t)^3\tan^{-1}(d+2r)/(s-t)+\\
&16(d+2r)^3(s-t)\tan^{-1}(s-t)\div\\
&(d+2r)-8(d+2r)(s-2t)^3\times\\
&\tan^{-1}(d+2r)/(s-2t)-8(d+2r)^3\times\\
&(s-2t)\tan^{-1}(s-2t)/(d+2r)+\\
&16(d+t)(s-t)^3\tan^{-1}(d+t)/(s-t)+\\
&16(d+t)^3(s-t)\tan^{-1}(s-t)/(d+t)-\\
&8(d+2t)(s-2t)^3\tan^{-1}(d+2t)\div\\
&(s-2t)-8(d+2t)^3(s-2t)\times\\
&\tan^{-1}(s-2t)/(d+2t)-16(d+r+t)\times\\
&(s-t)^3\tan^{-1}(d+r+t)/(s-t)-\\
&16(d+r+t)^3(s-t)\tan^{-1}(s-t)\div\\
&(d+r+t)+16(d+r+t)(s-2t)^3\times\\
&\tan^{-1}(d+r+t)/(s-2t)+\\
&16(d+r+t)^3(s-2t)\tan^{-1}(s-2t)\div\\
&\qquad(d+r+t)-8t^4(\pi+\log 2)\}\qquad(10)
\end{aligned}
$$

Corroboration of Equation 10

In other fashion each of the channels can be considered as comprised of three component conductors, each of rectangular cross section, and all having current in the same direction (Figure 3). Thus $m=n=3$.

By virtue of the geometric symmetries in Figure 3,

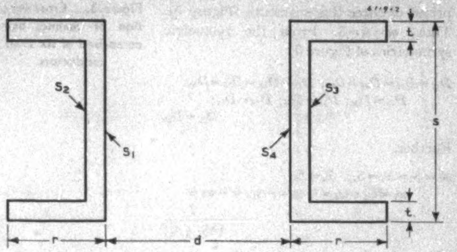
$$D_{11}=D_{33}=D_{44}=D_{66};\ D_{22}=D_{55};$$
$$D_{12}=D_{23}=D_{45}=D_{56};\ D_{13}=D_{46};$$
$$D_{16}=D_{34};\ D_{15}=D_{35}=D_{24}=D_{26};\ D_{14}=D_{36}$$

Further,

$$S_1=S_3=S_4=S_6;\ S_2=S_5;$$

$$w_1=w_2=w_3=-w_4=-w_5=-w_6=\frac{I}{A}$$

Figure 2. Cross section of channel bus comprised of four superimposed strap conductors



In consequence of these relationships, we have from equation 2

$$L=\left(-\frac{4}{A^2}\right)[2S_1^2(\log D_{11}+\log D_{13}-\log D_{14}-\log D_{16})+S_2^2(\log D_{22}-\log D_{25})+4S_1S_2(\log D_{12}-\log D_{15})]\qquad(11)$$

Corroboratively, if $S_1=S_3=S_4=S_6=0$ we have rectangular strap conductors of cross section $S_2=S_5$ and equation 11 reduces to the known formula,

$$L=4\log\frac{D_{25}}{D_{22}}$$

Determining the eight geometric mean distances in the right-hand member of equation 11 by use of equations 3–6—the explicit expressions are to be found in Appendix II—substituting appropriately in equation 11, and rearranging terms we obtain, as is to be expected, equation 10.

Although this verification most certainly ensures that the algebra involved in the derivation of the geometric mean distances given in Appendixes I and II and in the subsequent substitution of these expressions in equations 8 and 11 has been carried through correctly, certain additional checks are made easily.

1. If $r=0$, we have a bus comprised of strap conductors of width t and height s, spaced distance d between adjacent faces. Corroboratively, we obtain from equation 10 the known formula[9,10] for this case.
2. If $d=-r$, the conductor cross sections are located as in Figure 4; over the superimposed areas the current density is uniform and opposite; whence effectively we have the cross section of a bus comprised of strap conductors (cross-hatched) of width t and height $s-2t$, spaced distance $r-2t$ between adjacent faces. Corroboratively, we obtain from equation 10 the known formula[9,10] for this case.

Approximations to the General Formula

The general formula for the inductance comprises 75 terms: of these, 32 are logarithmic, 42 arctangentic, and 1 algebraic. This is a formidable number. But the work incidental to computation of the inductance of a specified bus can be carried through without difficulty provided in calculating one employs proper technique (judicious use of the trigonometric identity, $\tan^{-1}X+\tan^{-1}X^{-1}=\pi/2$; factoring, in so far as is possible, logarithmic arguments into prime numbers; and so forth) and suitable tools (an electric calculating machine; the 15-place Works Projects Administration tables[11] of natural logarithms and arc tangents having 0.0001 and 0.001 increment arguments). Admittedly, the task of computation is lengthy. In a sense, however, this is no drawback if, the final dimensions of the bus having been decided on, it is imperative to determine (as accurately as possible) the inductance for some specified purpose: say, to enable computation of the bus impedance for use in calculating voltage regulation or fault conditions. But in calculations leading to the choice of final bus dimensions, it is obviously desirable to have formulas that will enable rapid determination of the approximate inductances of the several designs under consideration.

A possible means of deriving such formulas is suggested by an earlier investigation of the inductances of busses comprised of hollow rectangular conductors.[12] It was found that for conductor thicknesses used in practice the inductance calculated from the formula taking conductor thickness into account and the inductance calculated from much simpler formulas derived by Dwight and Wang[5] on the basis that the conductor is infinitely thin—that is, that the cross section is comprised of line segments—differed only slightly. Although for reasons discussed in "Remarks on the Approximate Formula" such excellent agreement does not attend a similar treatment of channel conductors, it does afford formulas sufficiently satisfactory for preliminary calculations.

Consider, then, each conductor as comprised of three line segments (Figure 5). Thus, $m=n=3$. From the geometric symmetries of Figure 5,

$$D_{11}=D_{33}=D_{44}=D_{66};\ D_{12}=D_{23}=D_{45}=D_{56};\ D_{13}=D_{46};\ D_{14}=D_{36};\ D_{16}=D_{34};\ D_{22}=D_{55}$$

Further,

$$S_1=S_3=S_4=S_6;\ \ S_2=S_5;\ \ w_1=w_2=w_3=-w_4=-w_5=-w_6=\frac{I}{(2S_1+S_2)}$$

In virtue of these relationships, we have from equation 2

$$L=\left(\frac{-4}{B^2}\right)[2S_1^2(\log D_{11}+\log D_{13}-\log D_{14}-\log D_{16})+S_2^2(\log D_{22}-\log D_{25})+4S_1S_2(\log D_{12}-\log D_{24})] \quad (12)$$

wherein

$$B=(2S_1+S_2)=2x+y \quad (13)$$

Corroboratively, if $S_1=S_3=S_4=S_6=0$, we have infinitely thin strap conductors of linear cross section $S_2=S_5$, and equation 12 reduces to the known formula,

$$L=4\log\frac{D_{25}}{D_{22}}$$

Determining the eight geometric mean distances in the right-hand member of equation 12—the explicit expressions are to be found in Appendix III—and substituting appropriately in equation 12 yield

$$L=[2/(2x+y)^2]\{-4x^2\log x-6y^2\log y+4d^2\log d-4(d+x)^2\log(d+x)+2(x^2+2xy-y^2)\log(x^2+y^2)+2(d+2x)^2\log(d+2x)-4dy\log(d^2+y^2)-2[(d+x)^2-2(d+x)y-y^2]\times\log[(d+x)^2+y^2]+[(d+2x)^2-y^2]\times\log[(d+2x)^2+y^2]+4[x^2-2xy-y^2]\times\tan^{-1}x/y+4(d^2-y^2)\tan^{-1}d/y-4[(d+x)^2+2(d+x)y-y^2]\times\tan^{-1}(d+x)/y+4(d+2x)y\times\tan^{-1}(d+2x)/y+2\pi d(2x+y)\} \quad (14)$$

If $x=0$, we have a bus comprised of infinitely thin strap conductors of height y, spaced distance d. Corroboratively, we obtain from equation 14 the known formula[9,10] for this case. Again, if $d=-x$, the cross sections fall on one another; the superimposed flanges have equal and opposite currents densities; whence effectively we have a bus comprised of infinitely thin strap conductors of height y, spaced distance x. Corroboratively, we obtain from equation 14 the known formula[9,10] for this case.

The advantages of equation 14 over equation 10 in regard to computation are considerable. Whereas the general formula comprises 72 terms, this approximate formula comprises only 14: 9 logarithmic, 4 arctangentic, and 1 algebraic. Again, the coefficients and arguments of these terms are simpler than those of the corresponding terms of equation 10. In consequence, a calculation with this formula, if effected with the tools of computation mentioned above, requires but a short time. Unfortunately, for reasons discussed in "Remarks on the Approximate Formula," this approximate formula is, as mentioned above, not as accurate as that of the corresponding formula for hollow rectangular conductors.

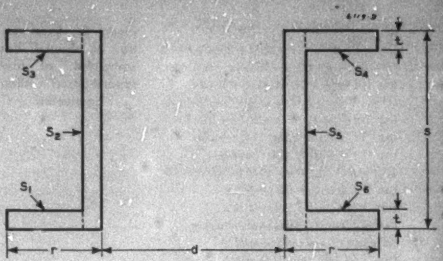
By means of the trigonometric identity, $\tan^{-1}X+\tan^{-1}X^{-1}=\pi/2$, the right-hand members of equations 10 and 14 can be expressed in various equivalent forms. However, in that the best use of this identity hinges on the numerical values associated with a given bus design—to avoid tedious interpolation with finite differences or computation with infinite series, this identity should be employed so that the maximum number of arctangent arguments are tabulated in the Works Projects Administration tables—there is little point in writing out any of the multitude of alternative forms.

Inasmuch as the right-hand members of equations 10 and 14 are dimensionless, we note that in substituting in these equations, r, s, t, x, y, and d can be measured in any like unit of length. But otherwise the units of all quantities mentioned to this point are in the absolute system: bus dimensions in centimeters; current I in abamperes; inductance L in abhenries per centimeter of bus length. Multiplying abhenries per centimeter by 3.048×10^{-5} yields millihenries per foot; multiplying by 11.48×10^{-6} yields 60-cycle reactance in ohms per foot.

Some Illustrative Examples

To illustrate use of the general formula and to gain knowledge of the accuracy to be expected of values calculated with the approximate formula, we consider the following problems:

Figure 3. Cross section of channel bus comprised of six strap conductors



EXAMPLE 1

To calculate the inductance of a bus comprised of (approximately) standard aluminum channels: web, $s=12$ inches; flange width, $r=3$ inches; thickness, $t=0.4$ inch; spaced back to back, $d=6$ inches (Figure 3):

Substituting these values in equation 10, collecting and rearranging numerical terms, obtaining values of the logarithmic and arctangentic terms from the Works Projects Administration tables, and effecting all calculations with an electric calculating machine, we obtain

$$L=2.58567_2 \text{ abhenries per centimeter of bus length} \quad (15)$$

Obviously, as by virtue of the assumptions underlying derivation of the general formula, it is unlikely that L represents the inductance of the hypothetical bus to more than plus or minus several in the second or third decimal place, the last several digits have no physical significance. They are given merely to emphasize that in so far as the numerical computation is concerned, the value of L is correct to this number of places. For all logarithmic and arctangentic terms were taken to nine decimal places; values stemming from consequent arithmetic manipulations were held to six places of decimals. In that the value given in the equation above stems from a final algebraic addition of some two score terms, the largest of which approximates 227,000 and the smallest 0.3, it is essential that some such degree of preciseness be held. In that the resulting value of L is significant to only two or three decimal places, it is not unlikely that study of the number of decimal places to be held in the various stages of computation would reveal that a smaller number than actually used would yet be satisfactory. Inasmuch, however, as 15-place tables were employed and the calculation was performed on a 10-digit electric calculating machine, the added labor is negligible and there is no uncertainty as to the effect of cumulative errors on the accuracy of first few digits of L.

EXAMPLE 2

To calculate the inductance of a bus comprised of channels having "external" dimensions and spacing as in example 1 ($s=12$ inches, $r=3$ inches, $d=6$ inches) but with thickness, $t=0.1$ inch.

Substituting appropriately in equation 10 and effecting computation as previously, we have

$L=2.53_8$ abhenries per centimeter of bus length (16)

Thus, as the thickness is decreased from 0.4 inch to 0.1 inch, the "external" dimensions and spacing of the channels being held constant, the inductance decreases slightly. Accordingly, it is tempting to conjecture that further reduction in thickness from 0.1 inch to indefinitely thin (line-segment) conductors will result in a correspondingly slight decrease in the inductance. That decrease in the thickness of the channels to line-segment channels does not result in a corresponding decrease in inductance is evidenced by the following example.

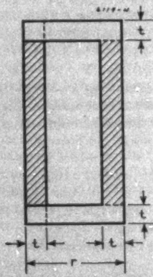


Figure 4. Cross section of rectangular strap bus comprised of two superimposed channels

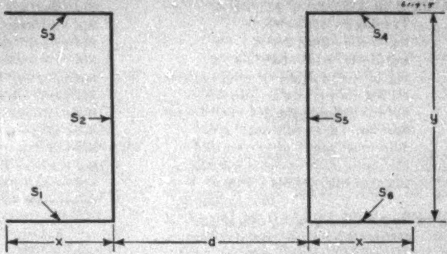
EXAMPLE 3

To calculate the inductance of a bus comprised of line-segment channels having "external" dimensions and spacing as in examples 1 and 2 ($y=12$ inches, $x=3$ inches, $d=6$ inches):

Substituting appropriately in equation 14 and effecting computation as before yield

$L_s=3.51_7$ abhenries per centimer of bus length (17)

Figure 5. Cross section of bus comprised of six infinitely-thin strap conductors



This value is considerably larger than that of equation 16.

Remarks on the Approximate Formula

Noting that the inductance of the line-segment bus is considerably larger than that of the original bus of example 1 gives rise to a question regarding the best use of the approximate formula to calculate the inductance of busses of finite thickness such as actually occur in practice. Thus, the "external" dimensions and spacing of the bus of example 1 are used in example 3. But if the "mean" or the "internal" dimensions and spacing were to be used, would not the resulting values of inductance be closer to those of example 1? To settle this point we calculate the inductances indicated.

EXAMPLE 4

To calculate the inductance of line-segment busses having (1) the "mean" and (2) the "internal" dimensions and spacing of example 1:

1. $x=2.8$ inches; $y=11.6$ inches; $d=6.4$ inches
2. $x=2.6$ inches; $y=11.2$ inches; $d=6.8$ inches

Substituting appropriately in equation 14 and effecting computation as before yield

$L=3.97_7$ abhenries per centimeter of bus length (18)

$L_i^m=4.38_7$ abhenries per centimeter of bus length (19)

We find, therefore, that the best approximation to the inductance of the bus of example 1 stems from use of the "external" dimensions and spacing.

From analysis of these four examples we conclude:

1. If the inductance of a bus is to be approximated by use of equation 14, the best result will stem from use of the "external" dimensions and spacing as in example 3.
2. Comparatively large changes in the thickness of the channels apparently result in slight changes in the inductance.
3. Though the approximate value will be somewhat larger than the actual inductance, it is yet of use in actual design.

For example, in the course of determining the best dimensions of a bus it is often necessary to determine whether or not the inductance of a number of proposed designs exceeds a certain allowed maximum—say that which specifies the maximum allowable voltage drop for a given load current. If of one or more of these designs it is found that the inductance calculated from the approximate equation is less than the allowed maximum, the exact value is even more so: whence the lengthy exact computation need not be undertaken.

Again, in much design work only an approximate knowledge of the inductance is needed; whence a value 20 or 30 per cent larger than the actual value is yet sufficiently accurate.

Appendix I

The geometric mean distances indicated in equation 8 follow from equation 3 by substitution therein of the following values of the parameters A_p, B_p, these being obtained by inspection of Figure 2.

$S_1^2 \log D_{11}$: $A_1=A_3=r$, $A_2=A_4=0$; $B_1=B_3=s$, $B_2=B_4=0$

$S_1^2 \log D_{14}$: $A_1=d+2r$, $A_2=A_4=d+r$, $A_3=d$; $B_1=B_3=s$, $B_2=B_4=0$

$S_2^2 \log D_{22}$: $A_1=A_3=r-t$, $A_2=A_4=0$; $B_1=B_3=s-2t$, $B_2=B_4=0$

$S_2^2 \log D_{23}$: $A_1=d+2r$, $A_2=A_4=d+r+t$, $A_3=d+2t$; $B_1=B_3=s-2t$, $B_2=B_4=0$

$S_1S_2 \log D_{12}$: $A_1=r-t$, $A_2=t$, $A_3=r$, $A_4=0$; $B_1=B_3=s-t$, $B_2=B_4=t$

$S_1S_2 \log D_{13}$: $A_1=d+2r$, $A_2=d+r$, $A_3=d+t$, $A_4=d+r+t$; $B_1=B_3=s-t$, $B_2=B_4=t$

Substituting these parameters in equation 3 yields:

$$S_1^2 \log D_{11}=-(25/12)S_1^2+(1/24)[4r^4\times \log r+4s^4 \log s-2(r^4-6r^2s^2+s^4)\times \log (r^2+s^2)+16rs^3 \tan^{-1} r/s+16r^3s \tan^{-1} s/r]$$

$$S_1^2 \log D_{14} = -(25/12)S_1^2 + (1/24) \times \{2d^4 \log d - 4(d+r)^4 \times \log (d+r) + 2(d+2r)^4 \times \log (d+2r) - (d^4 - 6d^2s^2 + s^4) \times \log (d^2+s^2) + 2[(d+r)^4 - 6(d+r)^2s^2 + s^4] \log [(d+r)^2+s^2] - [(d+2r)^4 - 6(d+2r)^2s^2 + s^4] \log [(d+2r)^2+s^2] + 8ds^3 \tan^{-1} d/s + 8d^3s \tan^{-1} s/d - 16(d+r)s^3 \tan^{-1} (d+r)/s - 16(d+r)^3s \tan^{-1} s/(d+r) + 8(d+2r)s^3 \times \tan^{-1} (d+2r)/s + 8(d+2r)^3s \times \tan^{-1} s/(d+2r)\}$$

$$S_2^2 \log D_{22} = -(25/12)S_2^2 + (1/24)\{4(r-t)^4 \times \log (r-t) + 4(s-2t)^4 \log (s-2t) - 2[(r-t)^4 - 6(r-t)^2(s-2t)^2 + (s-2t^4] \log [(r-t)^2+(s-2t)^2] + 16(r-t)(s-2t)^3 \tan^{-1}(r-t)/(s-2t) + 16(r-t)^3(s-2t) \times \tan^{-1} (s-2t)/(r-t)\}$$

$$S_2^2 \log D_{23} = -(25/12)S_2^2 + (1/24)\{2(d+2r)^4 \log (d+2r) + 2(d+2t)^4 \log (d+2t) - 4(d+r+t)^4 \log (d+r+t) - [(d+2r)^4 - 6(d+2r)^2(s-2t)^2 + (s-t)^4] \log [(d+2r)^2+(s-2t)^2] - [(d+2t)^4 - 6(d+2t)^2(s-2t)^2 + (s-2t)^4] \log [(d+2t)^2+(s-2t)^2] + 2[(d+r+t)^4 - 6(d+r+t)^2(s-2t)^2 + (s-2t)^4] \log [(d+r+t)^2+(s-2t)^2] + 8(d+2r)(s-2t)^3 \times \tan^{-1} (d+2r) \div (s-2t) + 8(d+2r)^3(s-2t) \tan^{-1} (s-2t)/(d+2r) + 8(d+2t)(s-2t)^3 \times \tan^{-1} (d+2t)/(s-2t) + 8(d+2t)^3 \times (s-2t) \times \tan^{-1} (s-2t)/(d+2t) - 16(d+r+t)(s-2t)^3 \tan^{-1}(d+r+t) \div (s-2t) - 16(d+r+t)^3(s-2t) \times \tan^{-1} (s-2t)/(d+r+t)\}$$

$$S_1S_2 \log D_{12} = -(25/12)S_1S_2 + (1/24) \times \{4t^4(\pi + \log 2) + 6t^4 \log t + 2(s-t)^4 \times \log (s-t) + (r^4 - 6r^2t^2 + t^4) \times \log (r^2+t^2) - [r^4 - 6r^2(s-t)^2 + (s-t)^4] \log [r^2+(s-t)^2] + [(r-t)^4 - 6(r-t)^2t^2 + t^4] \times \log [(r-t)^2+t^2] + [(s-t)^4 - 6(s-t)^2t^2 + t^4] \log [(s-t)^2+t^2] - [(r-t)^4 - 6(r-t)^2(s-t)^2 + (s-t)^4] \times \log [(r-t)^2+(s-t)^2] - 8rt^3 \times \tan^{-1} r/t - 8r^3t \tan^{-1} t/r - 8(r-t)t^3 \times \tan^{-1} (r-t)/t - 8(r-t)^3t \tan^{-1} t \div (r-t) - 8(s-t)^3t \tan^{-1} t/(s-t) - 8(s-t)t^3 \tan^{-1} (s-t)/t + 8r(s-t)^3 \times \tan^{-1} r/(s-t) + 8r^3(s-t) \tan^{-1} (s-t)/r + 8(r-t)(s-t)^3 \tan^{-1} (r-t) \div (s-t) + 8(r-t)^3(s-t) \tan^{-1} (s-t) \div (r-t)\}$$

$$S_1S_2 \log D_{13} = -(25/12)S_1S_2 + (1/24) \times \{[(d+t)^4 - 6(d+t)^2t^2 + t^4] \log [(d+t)^2+t^2] - [(d+r)^4 - 6(d+r)^2t^2 + t^4] \times \log [(d+r)^2+t^2] + [(d+2r)^4 - 6(d+2r)^2t^2 + t^4] \log [(d+2r)^2+t^2] - [(d+r+t)^4 - 6(d+r+t)^2t^2 + t^4] \times \log [(d+r+t)^2+t^2] - [(d+t)^4 - 6(d+t)^2(s-t)^2 + (s-t)^4] \log [(d+t)^2+(s-t)^2] + [(d+r)^4 - 6(d+r)^2 \times (s-t)^2 + (s-t)^4] \log [(d+r)^2+(s-t)^2] - [(d+2r)^4 - 6(d+2r) \times (s-t)^2 + (s-t)^4] \log [(d+2r)^2+(s-t)^2] + [(d+r+t)^4 - 6(d+r+t)^2 \times (s-t)^2 + (s-t)^4] \log [(d+r+t)^2+(s-t)^2] - 8(d+t)t^3 \tan^{-1} (d+t)/t - 8(d+t)^3t \tan^{-1} t/(d+t) + 8(d+r)t^3 \times \tan^{-1}(d+r)/t + 8(d+r)^3t \tan^{-1} t \div (d+r) - 8(d+2r)t^3 \tan^{-1} (d+2r)/t - 8(d+2r)^3t \tan^{-1} t/(d+2r) + 8(d+r+t)t^3 \tan^{-1} (d+r+t)/t + 8(d+r+t)^3t \tan^{-1} t/(d+r+t) + 8(d+t)(s-t)^3 \tan^{-1} (d+t)/(s-t) + 8(d+t)^3(s-t) \tan^{-1} (s-t)/(d+t) - 8(d+r)(s-t)^3 \tan^{-1} (d+r)/(s-t) - 8(d+r)^3(s-t) \tan^{-1} (s-t)/(d+r) + 8(d+2r)(s-t)^3 \tan^{-1}(d+2r)/(s-t) + 8(d+2r)^3(s-t) \tan^{-1}(s-t)/(d+2r) - 8(d+r+t)(s-t)^3 \tan^{-1} (d+r+t) \div (s-t) - 8(d+r+t)^3(s-t) \times \tan^{-1} (s-t)/(d+r+t)\}$$

Appendix II

The geometric mean distances indicated in equation 11 follow from equation 3 by substitution therein of the following values of the parameters A_p, B_q, these being obtained by inspection of Figure 3.

$S_1^2 \log D_{11}$: $A_1 = A_3 = r-t$, $A_2 = A_4 = 0$; $B_1 = B_3 = t$, $B_2 = B_4 = 0$

$S_1^2 \log D_{13}$: $A_1 = A_3 = r-t$, $A_2 = A_4 = 0$; $B_1 = s$, $B_2 = B_4 = s-t$, $B_3 = s-2t$

$S_1^2 \log D_{14}$: $A_1 = d+2r$, $A_2 = A_4 = d+r+t$, $A_3 = d+2t$; $B_1 = s$, $B_2 = B_4 = s-t$, $B_3 = s-2t$

$S_1^2 \log D_{16}$: $A_1 = d+2r$, $A_2 = A_4 = d+r+t$, $A_3 = d+2t$; $B_1 = B_3 = t$, $B_2 = B_4 = 0$

$S_2^2 \log D_{22}$: $A_1 = A_3 = t$, $A_2 = A_4 = 0$; $B_1 = B_3 = s$, $B_2 = B_4 = 0$

$S_2^2 \log D_{25}$: $A_1 = d+2t$, $A_2 = A_4 = d+t$, $A_3 = d$; $B_1 = B_3 = s$, $B_2 = B_4 = 0$

$S_1S_2 \log D_{12}$: $A_1 = r$, $A_2 = t$, $A_3 = 0$, $A_4 = r-t$; $B_1 = s$, $B_2 = s-t$, $B_3 = t$, $B_4 = 0$

$S_1S_2 \log D_{15}$: $A_1 = d+r+t$, $A_2 = d+r$, $A_3 = d+t$, $A_4 = d+2t$; $B_1 = s$, $B_2 = s-t$, $B_3 = t$, $B_4 = 0$

Substituting these parameters in equation 3 yields

$$S_1^2 \log D_{11} = -(25/12)S_1^2 + (1/24)\{4t^4 \times \log t + 4(r-t)^4 \log (r-t) - 2[(r-t)^4 - 6(r-t)^2t^2 + t^4] \log [(r-t)^2+t^2] + 16(r-t)t^3 \tan^{-1} (r-t)/t + 16(r-t)^3t^3 \tan^{-1} t/(r-t)\}$$

$$S_1^2 \log D_{13} = -(25/12)S_1^2 + (1/24)\{2s^4 \times \log s - 4(s-t)^4 \log (s-t) + 2(s-2t)^4 \times \log (s-2t) - [(r-t)^4 - 6(r-t)^2x^2 + s^4] \log [(r-t)^2+s^2] + 2[(r-t)^4 - 6(r-t)^2(s-t)^2 + (s-t)^4] \log [(r-t)^2+(s-t)^2] - [(r-t)^4 - 6(r-t)^2(s-2t)^2 + (s-2t)^4] \log [(r-t)^2+(s-2t)^2] + 8(r-t)s^3 \tan^{-1} (r-t)/s + 8(r-t)^3s \times \tan^{-1} s/(r-t) - 16(r-t)(s-t)^3 \times \tan^{-1} (r-t)/(s-t) - 16(r-t)^3 \times (s-t) \tan^{-1} (s-t)/(r-t) + 8(r-t)(s-2t)^3 \tan^{-1} (r-t)/(s-2t) + 8(r-t)^3(s-2t) \tan^{-1} (s-2t)/(r-t)\}$$

$$S_1^2 \log D_{14} = -(25/12)S_1^2 - (1/24)\{[(d+2r)^4 - 6(d+2r)^2s^2 + s^4] \log [(d+2r)^2+s^2]^{1/2} - [(d+2r)^4 - 6(d+2r)^2(s-t)^2 + (s-t)^4] \log [(d+2r)^2+(s-t)^2] + [(d+2r)^4 - 6(d+2r)^2(s-2t)^2 + (s-2t)^4] \log [(d+2r)^2+(s-2t)^2]^{1/2} + [(d+2t)^4 - 6(d+2t)^2s^2 + s^4] \times \log [(d+2t)^2+s^2]^{1/2} - [(d+2t)^4 - 6(d+2t)^2(s-t)^2 + (s-t)^4] \times \log [(d+2t)^2+(s-t)^2] + [(d+2t)^4 - 6(d+2t)^2(s-2t)^2 + (s-2t)^4] \log [(d+2t)^2+(s-2t)^2]^{1/2} - [(d+r+t)^4 - 6(d+r+t)^2s^2 + s^4] \log [(d+r+t)^2+s^2] + 2[(d+r+t)^4 - 6(d+r+t)^2(s-t)^2 + (s-t)^4] \times \log [(d+r+t)^2+(s-t)^2] - [(d+r+t)^4 - 6(d+r+t)^2(s-2t)^2 + (s-2t)^4] \times \log [(d+r+t)^2+(s-2t)^2] - 4(d+2r)s^3 \tan^{-1} (d+2r)/s - 4(d+2r)^3s \times \tan^{-1} s/(d+2r) + 8(d+2r)(s-t)^3 \times \tan^{-1} (d+2r)/(s-t) + 8(d+2r)^3 \times (s-t) \tan^{-1} (s-t)/(d+2r) - 4(d+2r)(s-2t)^3 \tan^{-1} (d+2r)/(s-2t) - 4(d+2r)^3(s-2t) \tan^{-1} (s-2t)/(d+2r) - 4(d+2t)s^3 \tan^{-1} (d+2t)/s - 4(d+2t)^3s \tan^{-1} s/(d+2t) + 8(d+2t)(s-t)^3 \tan^{-1} (d+2t)/(s-t) + 8(d+2t)^3(s-t) \tan^{-1} (s-t)/(d+2t) - 4(d+2t)(s-2t)^3 \tan^{-1} (d+2t) \div (s-2t) - 4(d+2t)^3(s-2t) \times \tan^{-1} (s-2t)/(d+2t) + 8(d+r+t)s^3 \tan^{-1} (d+r+t)/s + 8(d+r+t)^3s \tan^{-1} s/(d+r+t) - 16(d+r+t)(s-t)^3 \tan^{-1} (d+r+t) \div (s-t) - 16(d+r+t)^3(s-t) \times \tan^{-1} (s-t)/(d+r+t) + 8(d+r+t) \times (s-2t)^3 \tan^{-1} (d+r+t)/(s-2t) + 8(d+r+t)^3(s-2t) \tan^{-1} (s-2t) \div (d+r+t)\}$$

$$S_1^2 \log D_{16} = -(25/12)S_1^2 + (1/24) \times \{2(d+2r)^4 \log (d+2r) + 2(d+2t)^4 \times \log (d+2t) - 4(d+r+t)^4 \times \log (d+r+t) - [(d+2r)^4 - 6(d+2r)^2t^2 + t^4] \log [(d+2r)^2+t^2] - [(d+2t)^4 - 6(d+2t)^2t^2 + t^4] \times \log [(d+2t)^2+t^2] + 2[(d+r+t)^4 - 6(d+r+t)^2t^2 + t^4 \times \log [(d+r+t)^2+t^2] + 8(d+2r)t^3 \times \tan^{-1} (d+2r)/t + 8(d+2r)^3t \times \tan^{-1} t/(d+2r) + 8(d+2t)t^3 \times \tan^{-1} (d+2t)/t + 8(d+2t)^3t \times \tan^{-1} t/(d+2t) - 16(d+r+t)t^3 \times \tan^{-1} (d+r+t)/t - 16(d+r+t)^3t \times \tan^{-1} t/(d+r+t)\}$$

$$S_2^2 \log D_{22} = -(25/12)S_2^2 + (1/24)\{4s^4 \times \log s + 4t^4 \log t - 2(s^4 - 6s^2t^2 + t^4) \times \log (s^2+t^2) + 16st^3 \tan^{-1} s/t + 16ts^3 \tan^{-1} t/s\}$$

$$S_2^2 \log D_{25} = -(25/12)S_2^2 + (1/24)\{2d^4 \times \log d - 4(d+t)^4 \log (d+t) + 2(d+2t)^4 \log (d+2t) - (d^4 - d^2s^2 + s^4) \times \log (d^2+s^2) + 2[(d+t)^4 - 6(d+t)^2s^2 + s^4] \log [(d+t)^2+s^2] - [(d+2t)^4 - 6(d+2t)^2s^2 + s^2] \log [(d+2t)^2+s^2] + 8ds^3 \tan^{-1} d/s + 8d^3s \tan^{-1} s/d - 16(d+t)s^3 \tan^{-1} (d+t)/s - 16(d+t)^3s \tan^{-1} s/(d+t) + 8(d+2t)s^3 \times \tan^{-1} (d+2t)/s + 8(d+2t)^3s \times \tan^{-1} s/(d+2t)\}$$

$$S_1S_2 \log D_{12} = -(25/12)S_1S_2 + (1/24) \times \{2t^4(\pi + \log 2) + r^4 \log r - s^4 \log s + 2t^4 \log t - (r-t)^4 \log (r-t) + (s-t)^4 \times \log (s-t) - (r^4 - 6r^2s^2 + s^4) \times \log (r^2+s^2)^{1/2} - (r^4 - 6r^2t^2 + t^4) \times \log (r^2+t^2)^{1/2} + (s^4 - 6s^2t^2 + t^4) \times \log (s^2+t^2)^{1/2} + [(r-t)^4 - 6(r-t)^2s^2 + s^4] \times \log [(r-t)^2+s^2]^{1/2} + [(r-t)^4 - 6(r-t)^2t^2 + t^4] \times \log [(r-t)^2+t^2]^{1/2} - [(s-t)^4 -$$

$$\begin{aligned}
&6(s-t)^2t^2+t^4]\log[(s-t)^2+t^2]^{1/2}+\\
&[r^4-6r^2(s-t)^2+(s-t)^4]\times\\
&\log[r^2+(s-t)^2]^{1/2}-[(r-t)^4-\\
&6(r-t)^2(s-t)^2+(s-t)^4]\times\\
&\log[(r-t)^2+(s-t)^2]^{1/2}+\\
&4rs^3\tan^{-1}r/s+4r^3s\tan^{-1}s/r+\\
&4rt^3\tan^{-1}r/t+4r^3t\tan^{-1}t/r-\\
&4st^3\tan^{-1}s/t-4s^3t\tan^{-1}t/s-\\
&4(r-t)s^3\tan^{-1}(r-t)/s-4(r-t)^3s\times\\
&\tan^{-1}s/(r-t)-4(r-t)t^3\times\\
&\tan^{-1}(r-t)/t-4(r-t)^3t\times\\
&\tan^{-1}t/(r-t)+4(s-t)t^3\times\\
&\tan^{-1}(s-t)/t+4(s-t)^3t\times\\
&\tan^{-1}t/(s-t)-4r(s-t)^3\times\\
&\tan^{-1}r/(s-t)-4r^3(s-t)\times\\
&\tan^{-1}(s-t)/r+4(r-t)(s-t)^3\times\\
&\tan^{-1}(r-t)/(s-t)+4(r-t)^3\times\\
&\qquad(s-t)\tan^{-1}(s-t)/(r-t)\}
\end{aligned}$$

$$\begin{aligned}
S_1S_2\log D_{13}={}&-(25/12)S_1S_2-(1/24)\times\\
&\{(d+r)^4\log(d+r)-(d+t)^4\times\\
&\log(d+t)+(d+2t)^4\log(d+2t)-\\
&(d+t+r)^4\log(d+t+r)-\\
&[(d+r)^4-6(d+r)^2s^2+s^4]\times\\
&\log[(d+r)^2+s^2]^{1/2}-[(d+r)^4-\\
&6(d+r)^2t^2+t^4]\log[(d+r)^2+t^2]^{1/2}+\\
&[(d+t)^4-6(d+t)^2t^2+t^4]\times\\
&\log[(d+t)^2+t^2]^{1/2}-[(d+2t)^4-\\
&6(d+2t)^2t^2+t^4]\log[(d+2t)^2+t^2]^{1/2}+\\
&[(d+t)^4-6(d+t)^2s^2+s^4]\times\\
&\log[(d+t)^2+s^2]^{1/2}-[(d+2t)^4-\\
&6(d+2t)^2s^2+s^4]\log[(d+2t)^2+s^2]^{1/2}+\\
&[(d+r)^4-6(d+r)^2(s-t)^2+(s-t)^4]\times\\
&\log[(d+r)^2+(s-t)^2]^{1/2}-[(d+t)^4-\\
&6(d+t)^2(s-t)^2+(s-t)^4]\times\\
&\log[(d+t)^2+(s-t)^2]^{1/2}+[(d+2t)^4-\\
&6(d+2t)^2(s-t)^2+(s-t)^4]\times\\
&\log[(d+2t)^2+(s-t)^2]^{1/2}+\\
&[(d+r+t)^4-6(d+r+t)^2t^2+t^4]\times\\
&\log[(d+r+t)^2+t^2]^{1/2}+[(d+r+t)^4-\\
&6(d+r+t)^2s^2+s^4]\log[(d+r+t)^2+\\
&s^2]^{1/2}-[(d+r+t)^4-6(d+r+t)^2\\
&(s-t)^2+(s-t)^4]\times\log[(d+\\
&r+t)^2+(s-t)^2]^{1/2}+4(d+r)s^3\\
&\tan^{-1}(d+r)/s+4(d+r)^3s\times\\
&\tan^{-1}s/(d+r)+4(d+r)t^3\times\\
&\tan^{-1}(d+r)/t+4(d+r)^3t\times\\
&\tan^{-1}t/(d+r)-4(d+t)t^3\times\\
&\tan^{-1}(d+t)/t-4(d+t)^3t\times\\
&\tan^{-1}t/(d+t)+4(d+2t)t^3\times\\
&\tan^{-1}(d+2t)/t+4(d+2t)^3t\times\\
&\tan^{-1}t/(d+2t)-4(d+t)s^3\times\\
&\tan^{-1}(d+t)/s-4(d+t)^3s\times\\
&\tan^{-1}s/(d+t)+4(d+2t)s^3\times\\
&\tan^{-1}(d+2t)/s-4(d+r)(s-t)^3\times\\
&\tan^{-1}(d+r)/(s-t)-4(d+r)^3\times\\
&(s-t)\tan^{-1}(s-t)/(d+r)+4(d+t)\times\\
&(s-t)^3\tan^{-1}(d+t)/(s-t)+\\
&4(d+t)^3(s-t)\tan^{-1}(s-t)\div\\
&(d+t)+4(d+2t)^3s\tan^{-1}s/(d+2t)-\\
&4(d+2t)(s-t)^3\tan^{-1}(d+2t)\div\\
&(s-t)-4(d+2t)^3(s-t)\times\\
&\tan^{-1}(s-t)/(d+2t)-4(d+r+t)t^3\times\\
&\tan^{-1}(d+r+t)/t-4(d+r+t)^3t\times\\
&\tan^{-1}t/(d+r+t)-4(d+r+t)s^3\times\\
&\tan^{-1}(d+r+t)/s-4(d+r+t)^3s\times\\
&\tan^{-1}s/(d+r+t)+4(d+r+t)\times\\
&(s-t)^3\tan^{-1}(d+r+t)/(s-t)+\\
&4(d+r+t)^3(s-t)\tan^{-1}(s-t)\div\\
&\qquad(d+r+t)\}
\end{aligned}$$

Appendix III

The geometric mean distances indicated in equation 12 are given by equations 11 and 13 of reference 7. Appropriate substitution therein yields

$$S_1^2\log D_{11}=x^2(\log x-3/2)$$

$$\begin{aligned}
S_1^2\log D_{13}={}&(x^2-y^2)\log(x^2+y^2)^{1/2}+y^2\times\\
&\log y+2xy\tan^{-1}x/y-3x^2/2
\end{aligned}$$

$$\begin{aligned}
S_1^2\log D_{14}={}&{}^1/_2[(d+2x)^2-y^2]\log[(d+2x)^2+\\
&y^2]^{1/2}-[(d+x)^2-y^2]\log[(d+x)^2+\\
&y^2]^{1/2}+[(d^2-y^2)/2]\log(d^2+y^2)^{1/2}+\\
&(d+2x)y\tan^{-1}(d+2x)/y-\\
&2(d+x)y\tan^{-1}(d+x)/y+dy\times\\
&\tan^{-1}d/y-3x^2/2.
\end{aligned}$$

$$\begin{aligned}
S_1^2\log D_{16}={}&{}^1/_2\{(d+2x)^2\log(d+2x)-\\
&2(d+x)^2\log(d+x)+d^2\log d-3x^2\}
\end{aligned}$$

$$S_2^2\log D_{22}=y^2(\log y-3/2)$$

$$\begin{aligned}
S_2^2\log D_{25}={}&(y^2-d^2)\log(d^2+y^2)^{1/2}+d^2\times\\
&\log d+2dy\tan^{-1}y/d-3y^2/2
\end{aligned}$$

$$\begin{aligned}
S_1S_2\log D_{12}={}&{}^1/_2\{xy\log(x^2+y^2)+x^2\times\\
&\tan^{-1}y/x+y^2\tan^{-1}x/y-3xy\}
\end{aligned}$$

$$\begin{aligned}
S_1S_2\log D_{24}={}&{}^1/_2\{(d+x)y\log[(d+x)^2+\\
&y^2]-dy\log(d^2+y^2)+(d+x)^2\times\\
&\tan^{-1}y/(d+x)+y^2\tan^{-1}(d+x)/y-\\
&d^2\tan^{-1}y/d-y^2\tan^{-1}d/y-3xy\}
\end{aligned}$$

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Improvements in Performance of Hydroelectric Generating Units on the 2,000,000-Horsepower Saguenay System

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THE large number of similar high-capacity hydroelectric generating units in the power plants of the Saguenay system have afforded an unusual opportunity to effect improvements in efficiency, reliability, and output of such units. Well organized operating records, thorough engineering studies and experimentation, and farsighted management all have contributed to this result.

Following a brief review of the major features of the Saguenay system, criteria affecting rate of wastage of water-wheel-runner metal by pitting, and costs of repairs are discussed. The advantages of operating water wheels at best gate, and means of accomplishing this by gate position indicators, are noted. Increases in water-wheel efficiency and capacity effected through model and prototype tests are set forth. Methods employed to ensure maximum energy generation from available storage water are described. The effect of dirt on generator stator-winding temperature rise is investigated. Improved generator performance arising from better application of stator-winding insulation, superior stranding, transposition of stator coils, use of high-silicon stator iron, and better ventilating systems, is discussed at some length.

The Saguenay System

The Saguenay system, a 60-cycle system based on the Isle Maligne and Shipshaw hydroelectric developments, has been built to a capacity of 2,000,000 horsepower during the last quarter century. The Isle Maligne plant, when placed in service, was possibly the largest water-power development completed in one step. Isle Maligne has an installed capacity of 540,000 horsepower, and Shipshaw 1 (Chute-a-Caron), 300,000 horsepower. Shipshaw 2 plant has a capacity of 1,200,000 horsepower.

The Saguenay system is located in the valley of the Saguenay River and the broad basin of Lake St. John lies at its head. The watershed tributary to the Saguenay (Figure 1) is about 30,000 square miles in extent, and is richly endowed with water-power resources. Useful storage in Lake St. John and two other storage reservoirs, one completely artificial, now totals 400 billion cubic feet.

Early power developments in the area were located on the smaller and more easily harnessed tributaries of Lake St. John and the Saguenay River, rather than on the Saguenay itself. These plants were built principally to supply mills of the pulp and paper industry. There were 13 scattered plants with a total capacity of some 67,000 horsepower by 1925 when Isle Maligne was placed in service. It contains 12 45,000-horsepower units.

The transmission network of the Saguenay system came into being as the aluminum, pulp, and paper industries were attracted to the area by the presence of cheap power and raw materials, with the construction of the Isle Maligne development. It was expanded with the addition of Shipshaw 1 in 1931 and brought to its present stage, as shown by Figure 2, with the completion of Shipshaw 2 in 1943.

Although the complete Shipshaw power development had been planned when the river closure and spillways associated with Shipshaw 1 station were started, the depression of the '30s precluded construction, so it was not until the ever increasing need of aluminum for military aircraft had exhausted all quickly available power resources in Canada and the United States that actual construction of Shipshaw 2 was put in hand. The first unit was placed in service November 23, 1942, just 18 months and 8 days after the start of construction. Additional units were put into operation at a steady rate until December 24, 1943, when the twelfth and last unit went into service. Shipshaw 2 is one of the greatest single hydroelectric powerhouse in the world, capable of generating 1,200,000 horsepower at maximum capacity.

The addition of Shipshaw 2 to the Saguenay system necessitated careful planning to limit fault currents to values interruptible by available circuit breakers having a capacity of 2,500,000 kva at seven to eight cycles interrupting time. A high degree of reliability was required on account of the large d-c aluminum reduction load of 890,000 kw, and the necessity of maintaining synchronism with other power systems hundreds of miles away for power exchange for seasonal and daily load-factor equalization. A flexible system was also an imperative requirement to facilitate obtaining optimum efficiency in conversion of storage water during the winter period of low river flow.

A ring bus at generator voltage at Isle Maligne and a unit system at each Shipshaw plant, whereby one generator feeds a 154-kv transmission circuit with step-up and step-down transformers, directly supplying a block of the aluminum plant load substantially equal to its capacity, and with the surplus or deficiency handled over the 154-kv busses, provides reliability and flexibility. It also permits economy in usage of storage water. Fault currents are limited by splitting the 154-kv system into three subsystems which which can be paralleled through three 90,000-kva transformer banks onto a 13.2-kv bus at Arvida substation, or through three 56,250-kva transformer banks onto a 13.2-kv bus at Shipshaw 1.

Saguenay System Loads

Loads on the Saguenay system are almost entirely industrial, concentrated in large blocks, with very high daily load factors. Yearly system load factor was 74.7 per cent in 1939 and 78.8 per cent in 1945.

The aluminum industry's load in the Saguenay area reached 890,000 kw during the war, of which about 745,000 kw was converted to direct current by mercury-type rectifiers, most ignitrons, and an additional 88,000 kw was converted by synchronous converters. Local paper mills required 100,000 kw. At times, 285,000 kw have been delivered to and 204,000 kw received from a double-circuit line connecting the Saguenay system with the Shawinigan Water and Power Company's system in the St. Maurice Valley, 200 miles away.

During periods of high and normal flow in the Saguenay, and when aluminum production is below normal, secondary power is available for the generation of steam. Electric steam generators on the system

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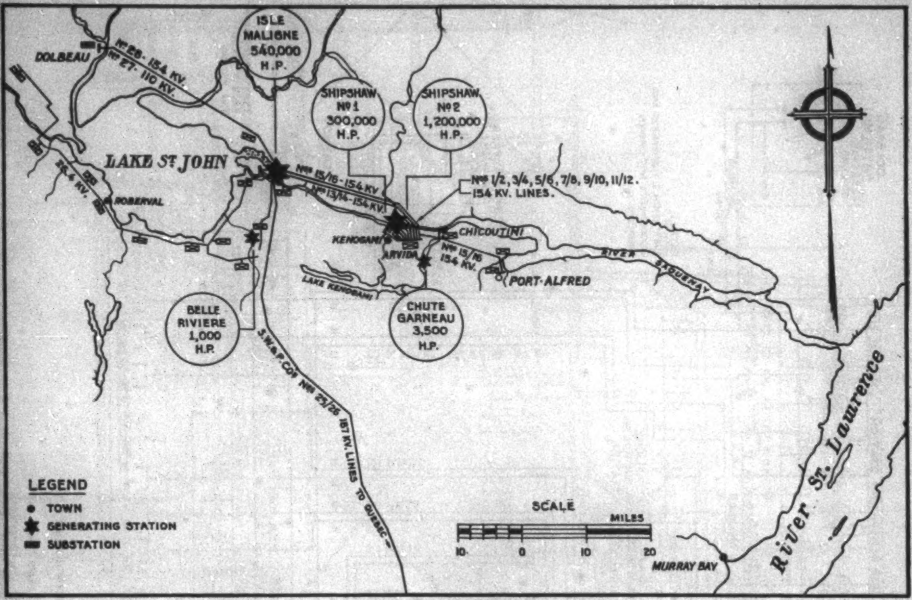

ISLE MALIGNE 540,000 H.P.
SHIPSHAW Nº 1 300,000 H.P.
SHIPSHAW Nº 2 1,200,000 H.P.
DOLBEAU
Nº 28 - 154 KV.
Nº 27 - 110 KV.
LAKE Sᵀ JOHN
26.4 KV.
ROBERVAL
Nºˢ 15/16 - 154 KV
Nºˢ 13/14 - 154 KV.
Nºˢ 1/2, 3/4, 5/6, 7/8, 9/10, 11/12.
154 KV. LINES.
KENOGAMI
ARVIDA
CHICOUTIMI
Nºˢ 15/16 154 KV.
PORT·ALFRED
RIVER SAGUENAY
LAKE KENOGAMI
BELLE RIVIERE 1,000 H.P.
CHUTE GARNEAU 3,500 H.P.
S.W.&P. COˢ Nºˢ 25/26 187 KV. LINES TO QUEBEC
River St. Lawrence
MURRAY BAY
LEGEND
TOWN
GENERATING STATION
SUBSTATION
SCALE
MILES
10
0
10
20

Figure 1. Map showing Saguenay system

have a total capacity of 360,000 kw. Primary power and occasionally a considerable amount of surplus power, for water storage purposes and for steam generation, is applied to the Shawinigan system.

Isle Maligne Generating Station

Isle Maligne generating station,[1] the first hydroelectric development on the Saguenay River, is located about nine miles below the outlet of Lake St. John. Owned and operated by Saguenay Power Company, Ltd., it contains 12 vertical Francis water wheels, each rated 45,000 horsepower, 112.5 rpm, 110 feet head, with direct-connected 35,000-kva 13.2-kv generators. A typical cross section of the plant is shown by Figure 3.

The generators are ventilated by an open system, cool air being taken from over the tailrace, through the wheel pits and generators to the generator room, whence it is discharged to atmosphere through the generator room windows and roof ventilators. During cold weather, part may be recirculated back through the generators by way of generator floor openings in order to maintain temperatures in the plant at comfortable levels.

Water-Wheel-Runner Pitting and Repairs

The water-wheel runners, measured at the middle of the guide case, have a diameter of 128 inches. A homologous runner, with 30.2 inch diameter, gave 93.3 per cent efficiency on a Holyoke test. The specific speed is 67, which the specific-speed experience curve indicates is in excess of the limiting value for freedom from pitting. Operating experience has confirmed this, considerable pitting having occurred.

The original runners were cast iron. Fatigue stresses in the vanes, where they join the crown plate, were increased by the removal of metal near the junction of the vanes and shroud ring through pitting. The net result was the gradual development of cracks in the vanes near the crown plate, starting from the discharge edges of the vanes and working outward. These cracks have been repaired by electric welding and mechanical clamps. All but one of the original 12 cast-iron runners are still in service. However, four cast-steel runners have been purchased. Three runners are always in the repair bay, which allows time for unhurried maintenance. Thirty-six hours' time is ample for exchanging runners.

Careful operating and inspection records were kept from the inception of operation. These included

1. Runner number.
2. Water-wheel number in which installed.
3. Date of inspection.
4. Number of hours in operation since last inspection.
5. Average operating load since last inspection.
6. Average operating gate opening since last inspection.
7. Average operating gross head since last inspection.
8. Average operating static draft head since last inspection.
9. Average effective draft head since last inspection.
10. Hours in operation with gate openings equal to or in excess of seven-tenths.
11. Hours in operation with load equal to or in excess of 30,000 kw.
12. Cracks.
13. Extent of pitting, plotted on standard forms for each vane, with contours showing area and depth, see Figure 4.
14. Cost of repairs.

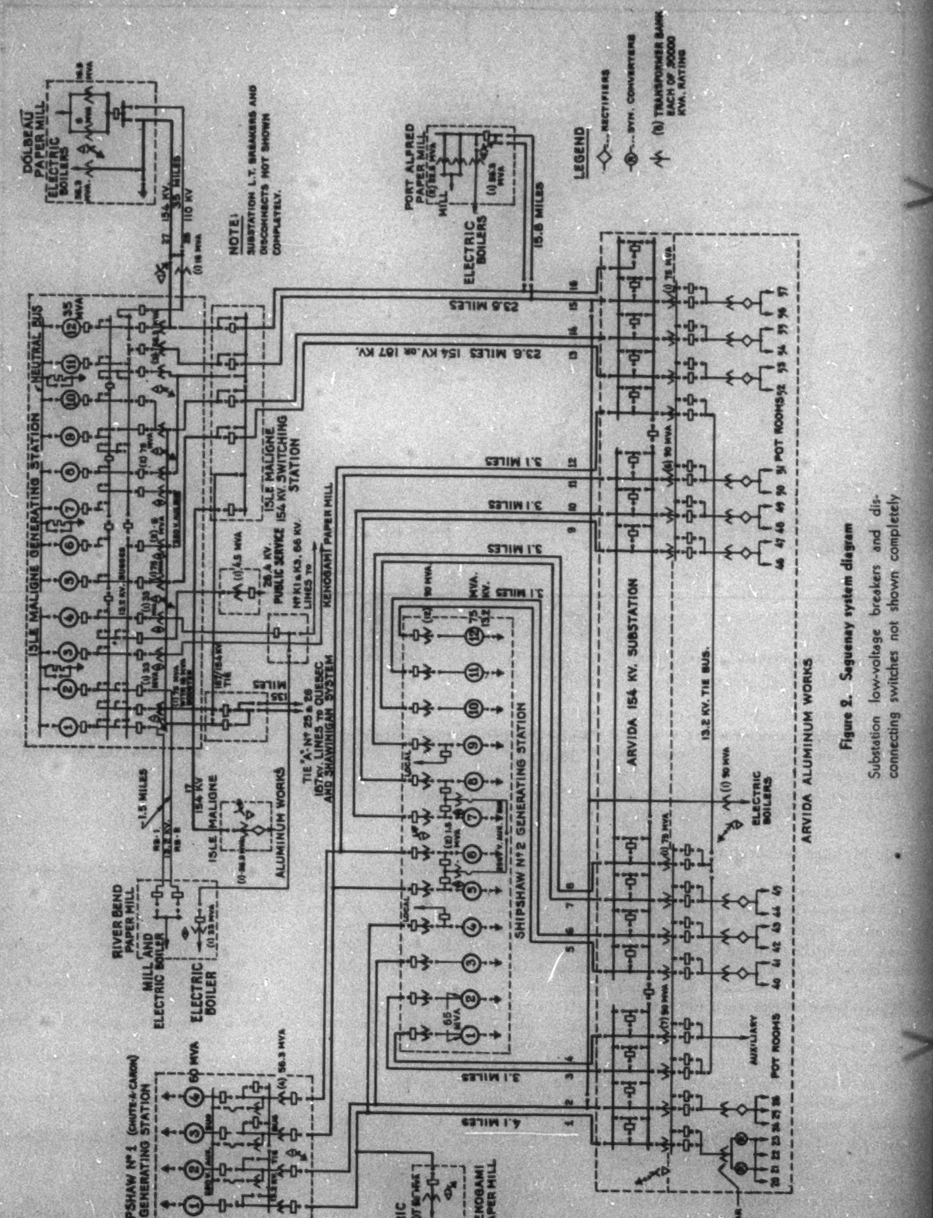

ISLE MALIGNE GENERATING STATION
NEUTRAL BUS
35 MVA
ISLE MALIGNE 154 KV. SWITCHING STATION
DOLBEAU PAPER MILL
ELECTRIC BOILERS
154 KV. 35 MILES
110 KV
NOTE: SUBSTATION L.T. BREAKERS AND DISCONNECTS NOT SHOWN COMPLETELY.
PORT ALFRED PAPER MILL
MILL
ELECTRIC BOILERS
15.8 MILES
LEGEND
RECTIFIERS
SYN. CONVERTERS
(6) TRANSFORMER BANK EACH OF 20000 KVA. RATING
23.6 MILES
23.6 MILES 154 KV. OR 187 KV.
3.1 MILES
26.4 KV. PUBLIC SERVICE
NO K1 & K3. 66 KV. LINES TO KENOGAMI PAPER MILL
TIE "A" NO 25 & 26 157 KV. LINES TO QUEBEC AND SHAWINIGAN SYSTEM
135 MILES
1.5 MILES
154 KV
RIVER BEND PAPER MILL
MILL AND ELECTRIC BOILER
ELECTRIC BOILER
ISLE MALIGNE ALUMINUM WORKS
SHIPSHAW NO 2 GENERATING STATION
65 MVA
75 MVA 13.2 KV.
LOCAL
ARVIDA 154 KV. SUBSTATION
13.2 KV. TIE BUS
ARVIDA ALUMINUM WORKS
POT ROOMS
AUXILIARY
ELECTRIC BOILERS
SHIPSHAW NO 1 (CHUTE-A-CARON) GENERATING STATION
60 MVA
56.3 MVA
4.1 MILES
KENOGAMI PAPER MILL
11 SIMILAR UNITS

Figure 2. Saguenay system diagram

Substation low-voltage breakers and disconnecting switches not shown completely

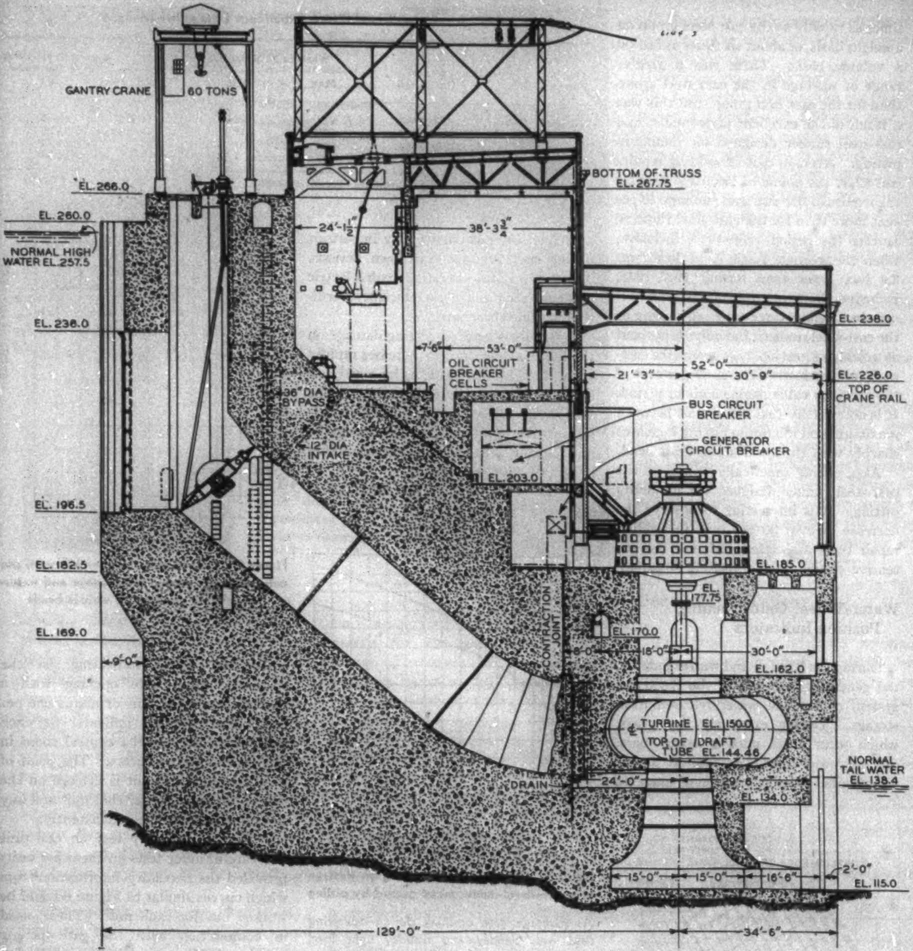

GANTRY CRANE
60 TONS
BOTTOM OF TRUSS
EL. 267.75
EL. 266.0
EL. 260.0
NORMAL HIGH WATER EL. 257.5
24'-1½"
38'-3¾"
EL. 238.0
EL. 236.0
7'-6"
53'-0"
OIL CIRCUIT BREAKER CELLS
52'-0"
21'-3"
30'-9"
EL. 226.0
TOP OF CRANE RAIL
36" DIA BYPASS
12" DIA INTAKE
BUS CIRCUIT BREAKER
GENERATOR CIRCUIT BREAKER
EL. 203.0
EL. 196.5
EL. 185.0
EL. 182.5
EL. 177.75
CONTRACTION JOINT
EL. 170.0
EL. 169.0
9'-0"
8'-0"
18'-0"
30'-0"
EL. 162.0
TURBINE EL. 150.0
TOP OF DRAFT TUBE EL. 144.46
NORMAL TAIL WATER EL. 138.4
DRAIN
24'-0"
9'-6"
EL. 134.0
15'-0"
15'-0"
16'-6"
2'-0"
EL. 115.0
129'-0"
34'-6"

Figure 3. Typical cross section through Isle Maligne power plant

15. Cost of welding cracks and pitted areas.

16. Electrodes used for item 15, in pounds.

17. Cost of special work.

18. Electrodes used for item 17, in pounds.

19. Runner outage for repairs, in hours.

20. Labor on repairs for item 14, in man-hours.

21. Cost per pound of electrodes used for repairing pitting and cracks.

The records show the wastage of metal due to pitting is closely proportional to hours of operation under either of two criteria during periods between repairs. These are:

1. Hours of operation at gate openings over seven-tenths.

2. Hours of operation over 30,000-kw load.

There is a definite variation between runners but for a given runner the foregoing is true.

Figure 5, derived from 4.5 years operation of a cast-steel runner, otherwise identical with the original cast-iron runners, illustrates the variation in wastage of metal due to pitting with gate opening. The wastage rapidly increases above seven-tenths gate. It may be noted seven-tenths gate is approximately best gate, at which maximum efficiency is obtained.

Salient data from the Isle Maligne experience on wastage of metal in water-wheel-runner vanes resulting from pitting, are embodied in Table I. During the period 1925 to 1938, inclusive, the group of cast-iron runners lost metal some 5.5

times as rapidly as the cast-steel group on a weight basis, or about six times as fast on a volume basis. There was a greater range of wastage in the cast-steel group than for the cast-iron group, but this was a result of the excellent performance of a cast-steel runner designed for minimum pitting. Average cost of welding repairs was $2.30 per pound of electrode used for the repairs to the cast-iron runners, 18 per cent more than for the cast-steel runners. During the period 1939–1944, inclusive, when the average static draft head was 3.4 feet lower than during 1925–1938, inclusive, the same group of cast-iron runners lost weight three times as rapidly as the cast-steel runners, but only 36 per cent as quickly as previously, whereas the cast-steel runners lost weight at 67 per cent of the previous value during a given period. It is noteworthy that the spread between maximum and minimum pitting increases sharply with the lower static draft head.

Also, pitting practically ceased on the cast-steel runner designed for minimum pitting. It is interesting to observe the decrease of over 30 per cent in welding repair costs associated with the less extensive pitting of the 1939–1944 period.

Table I. Water-Wheel Pitting Experience Data at Isle Maligne

Type of Runner	Period (Inclusive)	Wastage of Metal, Lb Per Year*			Cost of Repairs Per Lb of Electrode Used
		Max	Min	Avg	
Cast iron	1925–1938	744	144	411	$2.30
Cast steel	1925–1938	177	17	75	$1.95
Cast iron	1939–1944	509	61	149	$1.59
Cast steel	1939–1944	97	1.5	50	$1.33

* 8,760 hours of operation at gate openings over seven-tenths.

Water-Wheel Gate-Opening Position Indicators

During the late fall and winter months, the generating stations of the Saguenay system operate on water released from storage. During extremely dry years, which occur about once in seven years, storage water may be required during about nine months. Efficient usage of available water is particularly important during such storage drawdown periods. This necessitates operating hydroelectric units at their optimum efficiency, ordinarily termed best gate.

As previously noted, pitting damage on the Isle Maligne runners increases rapidly with time, during which the units are operated at gate openings in excess of the best gate. Consequently, normal operating procedure provides for operating the units at the gate openings corresponding to best gate for each unit.

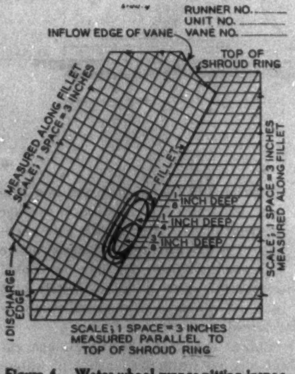

RUNNER NO.
UNIT NO.
VANE NO.
INFLOW EDGE OF VANE
TOP OF SHROUD RING
MEASURED ALONG FILLET SCALE; 1 SPACE = 3 INCHES
FILLET
1/8 INCH DEEP
1/4 INCH DEEP
3/8 INCH DEEP
SCALE; 1 SPACE = 3 INCHES MEASURED ALONG FILLET
DISCHARGE EDGE
SCALE; 1 SPACE = 3 INCHES MEASURED PARALLEL TO TOP OF SHROUD RING

Figure 4. Water-wheel-runner pitting inspection form

Typical pitting for cast-iron runner shown by contours

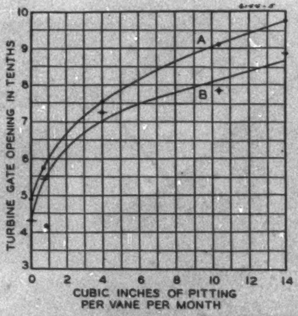


Figure 5. Effect of gate opening on wastage of water-wheel-runner metal caused by pitting

A—Average gate opening during operating time with Sunday and holiday light load periods deducted

B—Average gate opening during total operating time

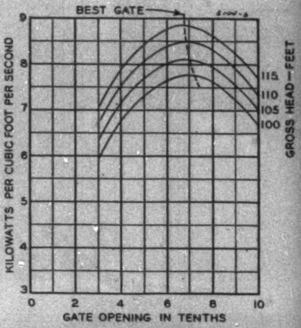


Figure 6. Relationship between kilowatts per cubic foot per second water usage and water-wheel gate opening for various heads

To facilitate this, a water-wheel gate-opening position indicator was devised. This consists of a Selsyn transmitter mechanically actuated by a rack on the water-wheel servomotor crosshead. A movement of the crosshead imparts a motion of rotation to the transmitter directly proportional to the guide-vane opening. This rotation of the transmitter causes unequal voltages to be set up in the circuit linking the transmitter and receiver, so that the receiver rotor moves to a position matching the transmitter rotor, indicating on the scale the exact gate opening with a maximum error of plus or minus one per cent. The receiving indicator for each unit is mounted in the control room in full view of the operators. The point of best gate for each unit is marked on the dial of the indicator for that unit and any deviation may be detected instantly.

A Gibson efficiency test on one unit and careful index tests on the other units provided the necessary information from which curves similar to Figure 6 could be worked out for each unit. These, used in conjunction with the gate-opening position indicators, afford an effective means of securing maximum energy from storage water and minimizing pitting damage because actual generation per shift can be compared with anticipated generation as derived from head and unit gate-opening records. These daily records are worked out by the shift operators and instill a spirit of competition, thus securing best results.

Effect of Dirt on Generator Temperature Rise

Hydroelectric generators with open ventilating systems, even though operating in a clean atmosphere, pick up an

Figure 7 (left). Dirt deposits at locations 1 and 2 on the stator of a 30,000-kva water-wheel generator operated for 5.67 years between thorough cleanings

Figure 8 (right). Dirt deposits at locations 3 and 4 on the stator of the same 30,000-kva water-wheel generator as shown in Figure 7

amazing amount of dirt which is deposited in the ventilating ducts and to a lesser extent on exposed portions of the windings. Combined with a slight amount of oil vapor which escapes from the guide bearings even with the most perfect seals, these deposits have sufficient coherence to adhere tightly and cannot be removed satisfactorily or completely with compressed air. These deposits interfere with the transfer of heat and reduce the amount of ventilating air. Cleaning practices vary but the most satisfactory appears to be about as follows.

When windings are dry and dusty, they can be blown out with dry compressed air and wiped down. The dirt in the ventilating ducts can be loosened up with rectangular-shaped pieces of fiber or suitable brushes and blown out. For moderately dirty windings with some oil, the coils can be wiped off with cloths dampened with varsol and dried by natural air currents. When windings are excessively dirty, with oil, they can be sprayed and flushed down with varsol, after which they can be blown out and wiped down. In such cases, the winding should be placed on dry out until the insulation resistance to ground returns to normal, as indicated by ohmmeter readings.

At Isle Maligne, it was found that the most satisfactory cleaning could be done when a unit was taken down for an exchange of runners. In one instance, a unit was operated 5.67 years between thorough cleanings of the stator ventilating ducts and winding. During this period, dirt deposits in the stator ducts built up as illustrated by Figure 7 and Figure 8.

In order to ascertain the effect of such deposits on the volume of cooling air passing through the ventilating ducts, a cylinder 36 inches long was fitted carefully over a typical area on the air gap side of the stator. At the end remote from the stator, dry compressed air under close control was admitted in a direction at right angles to the longitudinal axis of the cylinder, from an air receiver. Pressures were measured in a plane $2^{5}/_{8}$ inches away from the air gap face of the stator iron by a suitable probe and draft gauge. On the opposite side of the stator iron, and covering the same relative area, the air passing through the ducts was measured by a 4-inch diameter vane anemometer fitted in a $9^{1}/_{2}$-inch long cylinder. Suitable shoes and gaskets on the cylinders precluded any loss of air.

Tests at the four locations, indicated in Figures 7 and 8, gave closely similar results. Figure 9 is a plot of the variation in volume of air passing through the ducts in one square foot of stator surface, with respect to static pressure on the air gap face. The effect of the dirt deposits is readily apparent.

Careful anemometric tests, before and after thoroughly cleaning the generator, showed a reduction of 22.5 per cent in the total volume of cooling air passing through the generator, as a result of the 5.67 years' accumulation of dirt deposits.

The effect of these dirt deposits on the stator winding and iron temperatures is plotted in Figure 10, from which it will be noted the capacity of the generator was reduced by

1. 3,450 kva, or 11.5 per cent of the original 30,000 kva rating, for the same maximum winding temperature rise.
2. 5,800 kva or 19.3 per cent for the same average winding temperature rise.

Extensive studies on all elements of the generator ventilating circuit were made for the Isle Maligne units as a result of which worthwhile improvements were effected in the rotor baffling, as well as externally. This latter consisted of doubling the entry areas for the cooling air taken from over the tailrace by enlarging the turbine floor windows.

Generator Insulation

The original Isle Maligne generator windings are illustrated by the typical section in Figure 11. Winding failures began to occur after about five years

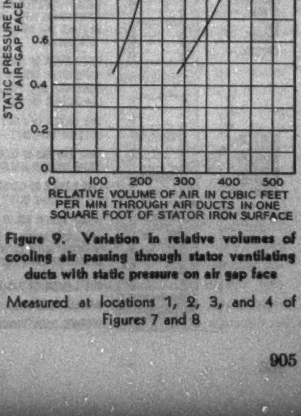


Figure 9. Variation in relative volumes of cooling air passing through stator ventilating ducts with static pressure on air gap face

Measured at locations 1, 2, 3, and 4 of Figures 7 and 8

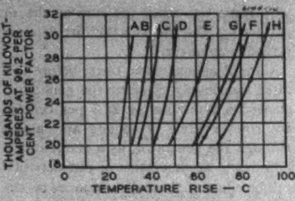


Figure 10. Effect of 5.67 years' accumulation of dirt deposits on stator iron and winding temperature rises of 30,000-kva generator

A—Average stator iron after cleaning
B—Maximum stator iron after cleaning
C—Average stator iron before cleaning
D—Maximum stator iron before cleaning
E—Average stator coil after cleaning
F—Maximum stator coil after cleaning
G—Average stator coil before cleaning
H—Maximum stator coil before cleaning

Table II. Water-Wheel Performance at Shipshaw 2 Generating Station

Item	Unit 8, With Original Design of Runner	Unit 10, With Runner Cut Back on Inlet Edges	Unit 9, With Runner Cut Back on Inlet and Discharge Edges
Output at best gate, horsepower	82,000	88,000	86,000
Output at full gate, horsepower	96,700	101,360	105,160
Efficiency at best gate, per cent	93.3	94.4	94.5
Efficiency at full gate, per cent	84.7	85.3	86.4
Water usage at best gate, cubic feet per second	3,727	3,953	3,859
Water usage at full gate, cubic feet per second	4,844	5,039	5,164

service and with increasing frequency. Cause of failures was found as follows:

1. Cotton and paper tape used to bind together the five turns which comprised a coil, before the application of the class B insulating jacket shrank and became brittle as a result of

2. High temperature accelerated by corona on the coils near the machine terminals where the voltage to ground was in excess of 7,000 rms volts.

3. Vibration, always potentially present, began to take place in the space provided by the shrinkage and the displacement of the embrittled insulation.

4. Eventually the mica strip between two turns was worn through or a strand worked around the edge of the strip, causing a short circuit of a turn, which heated the copper to the melting point. The insulating jacket next failed as a result of the heat and pressure of the molten copper.

An improved design of coil was worked out which almost eliminated the cellulose tape binding beneath the insulating jacket. It also contained 15 per cent more copper than the original coil with the same over-all thickness of insulation. This was accomplished by the use of less cellulosic carrier for the mica and a denser application, eliminating voids. Superior transposition of the winding was effected by the use of new series connectors. The new coil design, shown in Figure 11, also gave corona protection in the form of an Aquadag-treated asbestos outer wrapper. This design of stator winding was installed in all units and has given every indication of superior life.

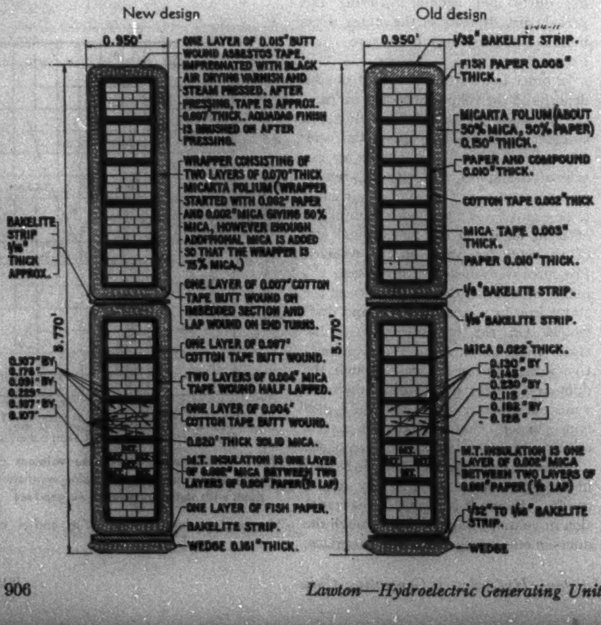
Isle Maligne Generator Losses

Together with an efficiency test by the Gibson method on the last water wheel installed, efficiency tests were made on

1. A generator with the original stator iron (2.2 silicon) and the original stator winding.

Figure 11. Typical section through slot portions of original and new Isle Maligne generator stator windings

The whole coil is impregnated with asphalt compound (crude oil specially processed, practically a synthetic asphalt)


New design
0.950'
ONE LAYER OF 0.015" BUTT WOUND ASBESTOS TAPE, IMPREGNATED WITH BLACK AIR DRYING VARNISH AND STEAM PRESSED. AFTER PRESSING, TAPE IS APPROX. 0.007" THICK. AQUADAG FINISH IS BRUSHED ON AFTER PRESSING.
WRAPPER CONSISTING OF TWO LAYERS OF 0.070" THICK MICARTA FOLIUM (WRAPPER STARTED WITH 0.002" PAPER AND 0.002" MICA GIVING 50% MICA, HOWEVER ENOUGH ADDITIONAL MICA IS ADDED SO THAT THE WRAPPER IS 75% MICA.)
BAKELITE STRIP 1/16" THICK APPROX.
ONE LAYER OF 0.007" COTTON TAPE BUTT WOUND ON IMBEDDED SECTION AND LAP WOUND ON END TURNS.
5.770'
ONE LAYER OF 0.007" COTTON TAPE BUTT WOUND.
0.307" BY 0.176"
0.091" BY 0.229"
0.307" BY 0.107"
TWO LAYERS OF 0.004" MICA TAPE WOUND HALF LAPPED.
ONE LAYER OF 0.004" COTTON TAPE BUTT WOUND.
0.020" THICK SOLID MICA.
M.T.
M.T. INSULATION IS ONE LAYER OF 0.002" MICA BETWEEN TWO LAYERS OF 0.001" PAPER (½ LAP)
ONE LAYER OF FISH PAPER.
BAKELITE STRIP.
WEDGE 0.161" THICK.
Old design
0.950'
1/32" BAKELITE STRIP.
FISH PAPER 0.008" THICK.
MICARTA FOLIUM (ABOUT 50% MICA, 50% PAPER) 0.150" THICK.
PAPER AND COMPOUND 0.010" THICK.
COTTON TAPE 0.002" THICK
MICA TAPE 0.003" THICK.
PAPER 0.010" THICK.
1/8" BAKELITE STRIP.
1/16" BAKELITE STRIP.
5.770'
MICA 0.022" THICK.
0.130" BY 0.145"
0.230" BY 0.115"
0.182" BY 0.128"
M.T.
M.T. INSULATION IS ONE LAYER OF 0.002" MICA BETWEEN TWO LAYERS OF 0.001" PAPER (½ LAP)
1/32" TO 1/16" BAKELITE STRIP.
WEDGE

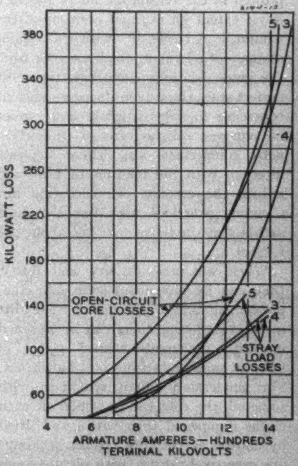


Figure 12. Improvement in core and stray-load losses of Isle Maligne generators resulting from use of higher silicon iron and new design of stator windings

Generator 3—Original iron and new stator winding
Generator 4—New iron and stator winding
Generator 5—Original iron and stator winding

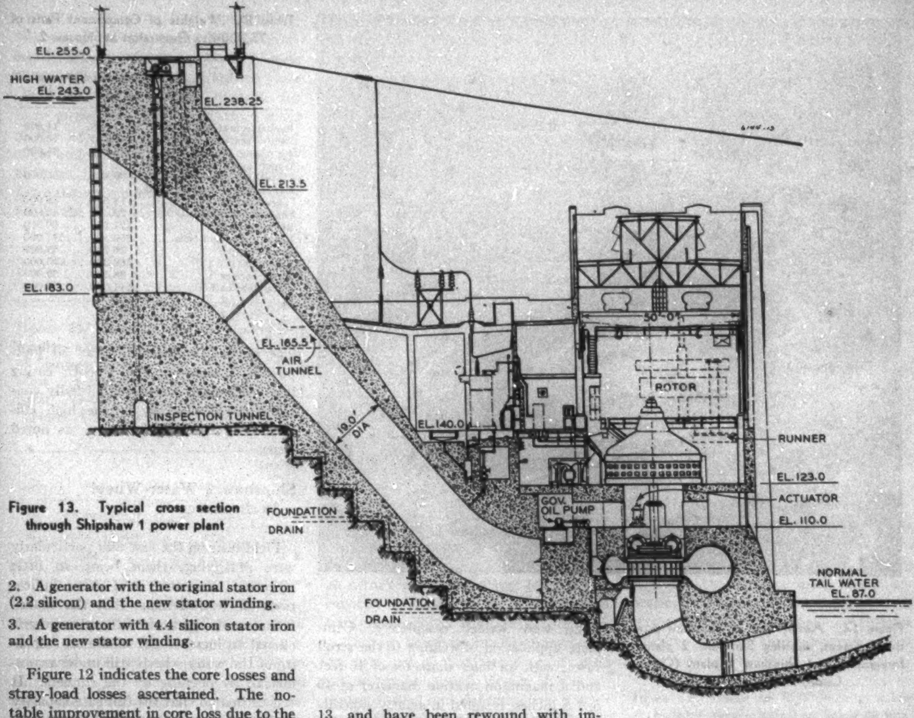


Figure 13. Typical cross section through Shipshaw 1 power plant

2. A generator with the original stator iron (2.2 silicon) and the new stator winding.
3. A generator with 4.4 silicon stator iron and the new stator winding.

Figure 12 indicates the core losses and stray-load losses ascertained. The notable improvement in core loss due to the use of the 4.4 silicon iron and in stray-load loss as a result of the new design of stator winding is significant of the improvements which frequently can be obtained with existing units by taking advantage of new materials, manufacturing technique, and design. The new windings and improved ventilation at Isle Maligne increased the output of each of the generators by 5,000 kva for the same temperature rise.

Shipshaw 1 Generating Station

Shipshaw 1 generating station, frequently called Chute-a-Caron, is located on the Saguenay River 20 miles below Isle Maligne. The first stage of the complete Shipshaw power development, owned by the Aluminum Company of Canada, Ltd., it contains four vertical Francis water-wheel units developing about 75,000 horsepower each, under 155 feet head, with 60,000-kva 13.2-kv 120-rpm generators. These generators have an open ventilating system, with cooling air being taken in from above the tailrace, as indicated by Figure 13, and have been rewound with improved coils similar to the new windings used at Isle Maligne.

Prior to the completion of the Shipshaw development by the construction of Shipshaw 2, the discharge from Isle Maligne plant was about 50 per cent in excess of the requirements at Shipshaw 1. Although the specific speed of 56 for the water wheels at the original rating of 65,000 horsepower was in excess of the value indicated by the experience curve for pitting, no pitting of any importance was encountered in operation. Also, it was known from test that the generators and transformers could carry safely at least 15 per cent overload.

Consequently, model tests were carried out to develop a runner capable of developing maximum output rather than maximum efficiency. These tests disclosed that a runner identical with the original design but with the diameter at the center line of the guide case reduced from 155 inches to 144.7 inches, could develop safely an additional 10,000 horsepower. New runners of the improved design were installed in all units, between November 1940 and October 1941, and operated with little pitting and efficiencies at wide gate openings better than the original runners.

Shipshaw 2 Generating Station

The Shipshaw 2 generating station, taking water from the same forebay as Shipshaw 1, is located about 1.5 miles down the Saguenay River from Shipshaw 1, as shown by the aerial view of Figure 14. Six 30-foot diameter shafts and tunnels each supply two single-runner, vertical, Francis-type water wheels originally specified to have 85,000 horsepower capacity at or near best gate, operating at 128.6 rpm. Eight units of one make are rated at 100,000 horsepower each and four of another make are rated at 85,000 horsepower each. Each turbine is directly connected to a three-phase generator having a specified capacity of 75,000 kva at 80-per cent power factor, 60-degree-centigrade maximum temperature rise, except two, which had been temporarily installed in an extension to Shipshaw 1 station, re-

Figure 14. Aerial view of Shipshaw power development, showing Shipshaw 2 plant in foreground and Shipshaw 1 plant (Chute-a-Caron) in left background

tained their 65,000 kva rating when relocated. Figure 15, a typical cross section through the headblock, water passages, and powerhouse, illustrates the general arrangement of the plant.

The scroll cases of the Shipshaw 2 water wheels,[2] including the connections to the cast-steel flanges of the speed rings, were welded completely. Complete application of welding to the scroll cases, with an inlet diameter of 16 feet and a maximum outside diameter of 49 feet 8 inches, resulted in greatly expediting field construction and very smooth casing outlines. All welding was done under extensometer and test-bar control. The former provided control of locked up stresses and eliminated the necessity of stress-relieving by controlled peening. The latter was taken care of by impact testing of sections cut daily from the actual welds of every operator.

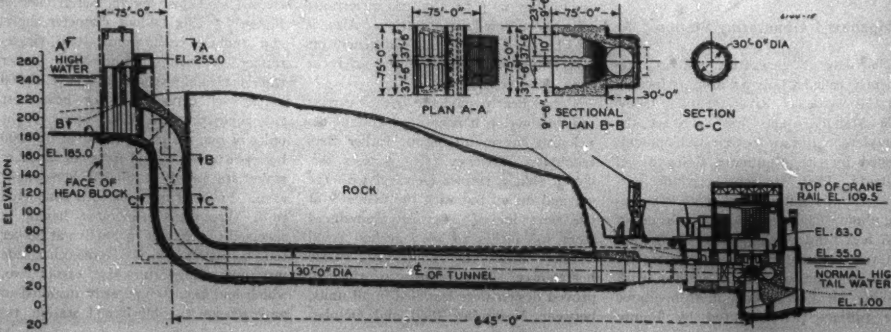
Table III. Weights of Component Parts of 75,000-Kva Generators at Shipsaw 2

	Generator A, Lb	Generator B, Lb
Exciter armature	12,800	11,500
Exciter stator	10,900	15,000
Air casing, louvres	35,000	35,000
Upper guide bearing and bracket	96,450	105,000
Lower guide bearing and bracket	18,600	35,000
Shaft, 35 in. diameter	76,000	63,000
Rotor, spider	121,850	60,000
Rotor, poles and coils	200,500	173,000
Rotor, copper	96,700	53,000
Rotor, net	766,300	623,000
Stator copper	33,500	39,000
Stator laminations	144,000	125,000

The general requirement was 30 foot-pounds (Charpy). The smooth casing lines with flush joints undoubtedly contributed substantially to the high efficiencies secured on field test, as noted later.

Shipshaw 2 Water-Wheel Performance

Field tests on the first unit particularly were gratifying, there being so little vibration, resonance, and other indications of overstressing, that the manufacturer of eight of the water wheels offered to increase the maximum capacity of the water wheels still under manufacture by opening up the runners. It was estimated that the full-gate capacity could be increased about 5,000 horsepower by cutting back the inlet edges of the runner vanes and a further 3,000 horsepower by cutting back on the discharge edges. It also was anticipated this in-



-75'-0"-
A
A
HIGH WATER
EL. 255.0
B
EL. 185.0
B
FACE OF HEAD BLOCK
C
C
ELEVATION
260
240
220
200
180
160
140
120
100
80
60
40
20
0
20
ROCK
30'-0" DIA
℄ OF TUNNEL
645'-0"
-75'-0"-
75'-0"
37'-6"
37'-6"
PLAN A-A
23'-0"
9'-6"
-75'-0"-
75'-0"
37'-6"
37'-6"
10'
9'-6"
30'-0"
SECTIONAL PLAN B-B
30'-0" DIA
SECTION C-C
TOP OF CRANE RAIL EL. 109.5
EL. 63.0
EL. 55.0
NORMAL HIGH TAIL WATER
EL. 1.00

Figure 15. Typical cross section through Shipshaw 2 power plant

crease in full-gate capacity would not be detrimental to the best-gate efficiency. No definite assurance could be given as to the effect on pitting, vibration, and resonance, but since the demand for more power was paramount, the runner of one unit was cut back on the inlet edges and that for a second unit was cut back on the inlet and discharge edges. After these were installed, efficiency tests using the Gibson method were made, with the results shown in Table II, from which it will be noted the expected increase in capacity was obtained. Following that, eight runners were trimmed on the inlet and discharge edges of the buckets. To date, after two years service, pitting has been very mild and efficiencies excellent.

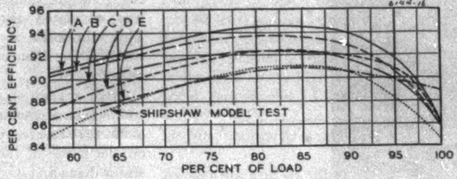


Figure 16. Comparative efficiencies of Shipshaw 2 and other outstanding water wheels in North America

Curve	Unit	Hp Rating	Ft. Head	Specific Speed, Rpm
A	Shipshaw unit 9	105,200	208	52
B	Niagara unit 21	88,000	213.5	39
C	Grand Coulee unit L7	115,500	330	33
D	Osage unit 3	37,000	90	78
E	Boulder unit 5	125,000	520	21

The relative efficiency of the Shipshaw and other outstanding water wheels in North America, all determined by the Gibson method, is illustrated by Figure 16. Note the striking correspondence between the actual field performance at Shipshaw and the model tests on a 14-inch 100-horsepower runner under a head of about 100 feet.

Generator Ventilation

In order to maintain the extremely close schedule under which Shipshaw 2 was built, it was necessary to utilize the manufacturing facilities of the only two Canadian manufacturers capable of building the generators. The order was divided equally, on the understanding all differences in electrical design and characteristics would be ironed out satisfactorily and that all units would be identical in outline and exterior dimensions.

A closed ventilating system was decided on, partly because of experience with the open ventilating systems at Isle Maligne and Shipshaw 1, and partly because the low temperature of the cooling medium, water from the forebay, would permit maximum utilization of the generator capacity at all seasons of the year. Moreover, the use of the closed ventilating system eliminated the necessity for otherwise lengthening the turbine shafts about 8 feet and providing an additional reinforced concrete floor over the space occupied by the main units to furnish necessary duct capacity. Actually, the whole structural layout was simplified and the cost decreased appreciably. Moreover, the lower, controllable ambient temperature permitted an actual increase of about 100,000 kva capacity for the station for the guaranteed temperature rise of 60 degrees centigrade.

The air housings for the generators[3] are arranged so that two doors at each corner can be opened, closing off the adjacent coolers and deflecting the hot air to the generator room for heating in cold weather. In the United States, it is usual practice to seal off the generator, but at Shipshaw this is accomplished by doors in the entrances to the wheel pits. The air from the generator rotor and stator is circulated through the water-cooled heat exchangers, whence it is discharged to the water-wheel pit and through the rotor to the stator.

The advantage of this method is the ease of access to the lower end turns of the stator winding, or the rotor guide bearing, without having to remove the sealing pans normally fixed to the generator lower bracket. In the case of the generators built by manufacturer *B*, the lower end turns are readily accessible in event of fire, without having to remove the baffles. Under actual conditions, the units built by manufacturer *A* take 135,000 cubic feet of cooling air per minute, whereas the others take 200,000 cubic feet per minute. These units have excess thermal capacity, and in all probability about 150,000 cubic feet of air per minute would have met requirements.

The rotating parts of the generator and turbine are carried on a single thrust bearing with one turbine and two generator guide bearings. Two types of thrust bearings are used, Kingsbury bearings and spring-thrust bearings. All thrust bearings, except one, are tin-base babbitt. On one thrust bearing, owing

Figure 17. Stator of 75,000-kva generator built by manufacturer *B* for Shipshaw 2

Note the curved ribs or spacer plates

Figure 18. View of rotor and stator of 75,000-kva generator built by manufacturer *B*

The air coolers can be seen at the corners of the housing

to the difficulty in obtaining tin, it was decided to try an arsenical lead babbitt which so far has proved satisfactory.

Manufacturer *A* used the conventional generator design with spider-type upper bracket and semidepressed oil pot around which the air casing was built and the coolers installed. The rotor is of standard design with cast-spider and conventional baffles. Manufacturer *B* employed a stator design which resulted in an open machine providing free circulation of air, and by shaping the spacer plates tended to control the direction of air flow towards the coolers. Note Figure 17. He used a depressed bridge-type girder upper-bracket arrangement with fully depressed thrust-bearing oil pot. A totally fabricated rotor was used, with the sections built up by welding steel plate. By means of specially designed slot wedges in the stator and special fan blades on the rotor, air flow through the stator is controlled without the use of baffles, leaving a very open machine on which the end turns, both upper and lower, are readily accessible. See Figure 18. This illustration shows the coolers at the corners of the housing.

The two different designs resulted in considerable difference in weight of materials, as shown in Table III. Table IV illustrates the comparative performance of the two types of generators.

In the generators built by manufacturer *B*, the greater part of the cooling air is supplied to the stator ducts through ducts in the rim of the rotor feeding the interpolar spaces, and then directly into the stator ducts, instead of the conventional manner wherein fans on the ends of the rotor blow air into the spaces between the poles. Thus the air entering the stator ducts is at the same temperature throughout the core, resulting in maximum and average temperatures closely alike. Scoop-type fans at the ends of the rotor ventilate the end turns and prevent recirculation of air. Actually, about 50 per cent more air than required to limit its temperature rise to the 18-degree-centigrade value desired, was circulated through the machine. This obviously resulted in a greater windage loss than otherwise would have been the case.

Table IV. Comparison of Shipshaw 2 *A* and *B* Generators Operating Under Full Load at Unity Power Factor

	Generator A	Generator B
Kilovolt-amperes or kilowatts	75,000	75,000
Stator current, amperes	3,280	3,280
Field current, amperes	740	855
Stator I^2R, kilowatts	244	199
Field I^2R, kilowatts	110	125
Open-circuit core loss, kilowatts	290	465
Short-circuit core loss, kilowatts	190	220
Windage and friction, kilowatts	825	555
Total losses, kilowatts	1,159	1,564
Output, kilowatts	75,000	75,000
Input, kilowatts	76,159	76,564
Efficiency, per cent	98.48	97.98

After a year of service, during the completion of construction of the plant, about 6.5 cubic feet of dirt were deposited on the inside of the rotor rim, but the stator ducts and duct entrances remained clean.

The *B* machines are designed for a different method of replacement of a few stator coils than the conventional method of dismantling the unit and removing the rotor.

The field coils and poles are arranged so they can be disconnected readily at the top. The pole keys are pulled out by the crane, and the rotor poles lifted out by the same means. The stator coil, or coils, are then removed by heating the adjacent group of coils electrically to soften the insulation, removing the slot wedges, and pulling the group out into the gap left by the withdrawn poles, and then lifting out the damaged coil or coils. New coils are installed by reversing this process. The working time required to make the replacement of a coil which failed was approximately 56 hours.

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The Electrical Performance of Ceramic Dielectrics at Elevated Temperatures

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Synopsis: Some recent developments have emphasized the need for information concerning the temperature at which dielectrics can be operated without electrical breakdown. This paper deals primarily with porcelain, but some data are included on the performance of steatite, glass, and zircon compositions. Data are presented on the variation in resistivity of the materials with varying temperature and also with varying voltage gradient. From the temperature–resistivity curves a method is developed for approximating the combination of temperature and voltage gradient which will result in electrical failure. Supporting data in the form of breakdown tests, made at various temperatures, are included. These tests include practical insulator forms as well as prepared test specimens.

IT IS WELL KNOWN that the electrical resistance of most of the commonly used insulating materials diminishes with increasing temperature. The actual performance of ceramic insulating materials at temperatures above the normal atmospheric level, and the factors affecting the ultimate breakdown at these temperatures, have not been the subject of much investigation. Recent requirements for insulators to operate at high temperatures and high voltages have indicated the need for such information.

There have been several insulator failures in precipitator applications which are directly attributable to the combination of high voltage and high temperature operation. Perhaps even more important is the development of certain new industrial processes which require that insulators be used at high temperatures. This investigation was undertaken for the purpose of obtaining fundamental design information on the high temperature electrical performance of ceramic materials and to develop suitable special materials for such applications.

Apparatus and Method

It was necessary to provide a suitable test chamber in which the specimens could be mounted and held at any desired temperature while voltage was applied. Accordingly, a thermally insulated box or furnace was built. The electrical heating elements in the back of the box were separated from the test chamber by a 1-inch thick thermally insulating partition in order to prevent direct heating of the specimens by radiant heat. The heated air was circulated by means of a fan through ports at each end of the test chamber. The fan also served to maintain air turbulence in the test chamber and thus prevent stratification. Temperature measurements were made by commercial gas thermometer, and automatic controls were provided for holding any desired temperature between minus 100 and plus 1,000 degrees Fahrenheit. With this arrangement temperatures could be held within five degrees Fahrenheit of the desired level. Provision also was made for removing the heating elements from the back of the box and substituting racks of dry ice in order to obtain low temperatures.

Most of the tests were made with direct voltage in order to eliminate the difficulty of separating conduction currents from charging currents. The voltage source was a full-wave high-voltage kenetron rectifier with suitable capacitors for the elimination of the 60-cycle pulse.

The direct voltages were measured by means of a suitable known resistance connected in series with a microammeter.

Most of the test specimens were prepared in the form of plates of various thicknesses. Some specimens were made from standard suspension insulator shells using the head section, ground down to the desired thickness, as the test area.

Electrodes of circular shapes were applied to the opposite sides of each test specimen. Guard rings also were provided to eliminate the effect of surface leakage and corona and to preserve a more nearly uniform field through the test zone. The electrodes and guard rings were of silver ceramicly bonded to the surface of the dielectric to avoid entrapped air between the dielectric and its electrode.

Resistance measurements were made by connecting one electrode to the voltage source and grounding the opposite electrode through a microammeter or milliammeter of proper scale range for the current involved. Suitable selector switches were provided for switching quickly from one instrument to another. The guard ring connections were shunted around the ammeters to ground. Resistance was determined from the ratio of voltage to current.

Before each measurement was made, the specimen was allowed to stand in the test chamber at zero voltage and constant temperature for a sufficient length of time to be sure that it was heated uniformly throughout its volume.

Materials

The greater part of this investigation is concerned with porcelain, although some data have been obtained on two kinds of glass, a steatite composition and a zircon composition.

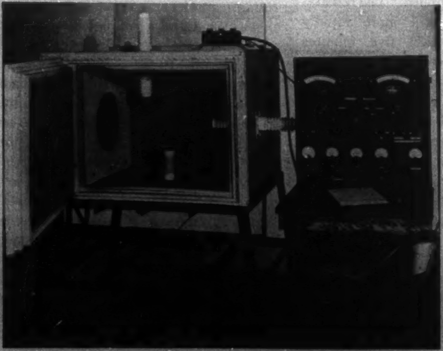

Figure 1. Test chamber and operating panel with specimen mounted in position

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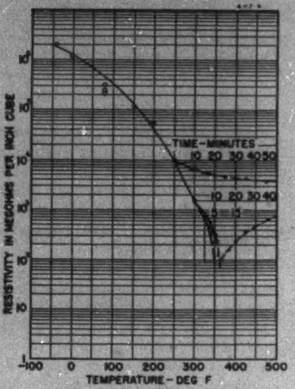

Figure 2. **Temperature versus resistivity, wet process electrical porcelain**

Voltage gradient 40 volts per mil direct current, thickness 0.8 inch, electrode area 30 square inches

These are all generic names, and there is considerable latitude in composition and in physical structure within each category.

The porcelain specimens involved all were made of wet process feldspathic porcelain of the type commonly known as electrical porcelain. Samples of such porcelain as made by three different manufacturers were used.

The steatite specimens were a magnesia-alumina-silicate composition maturing at a firing temperature of about 2,250 degrees Fahrenheit. The zircon specimens were composed basically of zirconia silicate with a calcium-zirconium-silicate flux compounded to mature at 2,250 degrees Fahrenheit. This composition was developed particularly for high temperature applications.

Two types of glass specimens were included: a borosilicate glass of the type used in the manufacture of high voltage insulators, and a lead glass also sometimes used in certain classes of insulators.

Discussion of Data—Porcelain

Many measurements were made on a variety of specimens of varying shapes and thicknesses, at temperatures ranging from minus 70 to plus 900 degrees Fahrenheit and at voltage gradients up to 700 volts per mil. A gratifying degree of consistency in results was obtained. Typical data illustrating the results obtained are presented herein. As an example, the curve shown in Figure 2 is typical of the measurement made to determine the variation of resistivity with temperature and to determine the temperature at which final breakdown would occur. In plotting the curve, the measured total resistances have been reduced to resistivity in megohms per inch cube. It should be noted that as the temperature was increased the resistivity steadily declined but remained stable (within the limits of measurement) at each point up to 250 degrees Fahrenheit. At 250 degrees Fahrenheit, on this particular specimen, it was found that with constant voltage the resistivity declined slowly with time, finally becoming stable at a lower level after about 45 minutes. This change in resistivity with time is indicated in Figure 2 by superimposing a time axis on the curve. The variation in resistivity with time, for a constant ambient temperature corresponding to that at zero time, is indicated by a broken line. This decline in resistance with time undoubtedly is caused by the internal heating of the specimen above the surrounding air because of the I^2R loss. It represents an average internal temperature rise of about 20 to 25 degrees Fahrenheit.

When the test chamber temperature was raised to 300 degrees Fahrenheit and held constant, the resistivity, with constant applied voltage, diminished with increasing rapidity until after 12 minutes the specimen no longer would support voltage and, for all practical purposes, had failed electrically. At this point voltage was removed, but the test chamber still was held at a constant temperature of 300 degrees Fahrenheit. At intervals of five minutes the same voltage was reapplied momentarily, and readings quickly were taken. It should be noted that as the specimen cooled internally the resistivity recovered along a typical cooling curve.

The test chamber temperature then was raised to 325 degrees Fahrenheit and the same voltage reapplied. The resistivity again declined still more rapidly with time, and within five minutes reached the point where voltage could not be supported. Voltage then was removed, and the piece again was allowed to recover. The test chamber temperature was raised to 350 degrees Fahrenheit and a momentary reading taken as indicated on the curve. The test chamber was allowed to cool back to room temperature, and the resistivity again was measured checking the value determined at the start of the test.

On this particular specimen, voltage was removed in each case as soon as a runaway condition was observed as indicated by current increasing more and more rapidly. On some other specimens, voltage was not removed promptly, and power was delivered to the piece for several seconds after the runaway phenomenon was observed. In such cases the dielectric was cracked or a fused path developed through it resulting in a permanent failure.

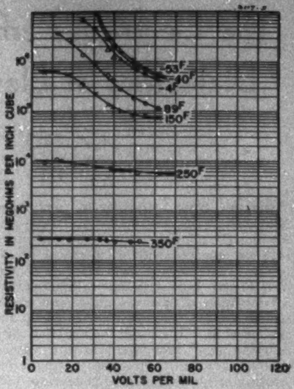

Figure 3. **Voltage gradient versus resistivity, wet process electrical porcelain**

Thickness 0.8 inch, electrode area 30 square inches

The curve shown in Figure 2 illustrates the variation of resistivity with temperature for one particular porcelain specimen at a constant voltage gradient. At any temperature level, the resistivity may vary also with changing voltage gradient. This effect was studied on a number of porcelain specimens, and Figure 3 shows a typical plot of the data obtained on one particular specimen—the same specimen appearing in Figure 2. It should be noted that at the lower temperatures resistivity diminishes rather rapidly with increasing voltage gradient, but at temperatures above 250 degrees Fahrenheit the resistivity becomes nearly independent of voltage gradient.

The results of all measurements made on porcelain to determine the variation of resistivity with temperature were plotted on one composite curve (Figure 4). At the higher temperature it was necessary to make all tests at relatively low voltage gradients in order to obtain stable measurements. Thus at 900 degrees Fahrenheit the measurements were made on a specimen approximately 0.8 inch thick with voltage applied from a 45-volt battery equivalent to about 0.05 volt per mil. At these higher temperatures the resistivity changes but little with changing voltage gradient so this limitation does not impair the usefulness of the information contained in Figure 4.

An examination of Figure 4 shows that data conform with reasonable accuracy to a single curve for all temperatures above about 250 or 300 degrees Fahrenheit. At lower temperatures the composite data spread widely as indicated by the envelope shown. This spread of the data at the lower temperatures is caused primarily by two factors:

1. The influence of voltage gradient on resistivity at lower temperature.
2. The influence of specimen thickness.

The influence of voltage gradient already has been illustrated. In testing specimens of different thicknesses, it was observed that at low temperatures the apparent resistivity was lower for thick specimens than for thin ones. The influence of thickness on resistivity diminished with increasing temperature so that for temperatures above about 250 or 300 degrees Fahrenheit the apparent resistivity was relatively independent of thickness. The apparent effect of thickness on resistivity is believed to be caused by the variation of the true voltage gradient from the average voltage gradient. As the specimen thickness is increased it becomes more and more difficult to maintain a truly uniform voltage gradient through it.

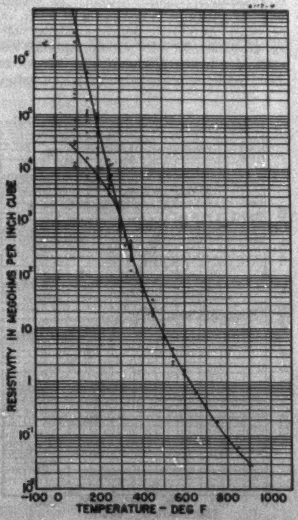

Figure 4. Composite curve, temperature versus resistivity of wet process electrical porcelain

It has been pointed out that porcelain made by three different manufacturers was used in the investigation. No significant difference was found among the three porcelains. Figure 5 shows a typical comparison of dimensionally identical pieces as made by two manufacturers. These two pieces were tested simultaneously. The change in resistivity with time (because of internal heating and cooling) is shown on the curve in the same manner as in Figure 2. At 300 degrees Fahrenheit, voltage was applied for ten minutes at which time the resistivity had become stable. Voltage then was removed and the specimen allowed to cool back to the test chamber temperature. Voltage then was reapplied, and a duplicate heating curve was obtained. At 325 degrees both specimens drifted toward breakdown. Specimen *B* actually refused to support voltage, but specimen *A* was well on the way to the same condition and doubtless would have failed in another minute or two had the test been continued. It should be noted that at 325 degrees both of the resistivity drift curves pass through a definite flex point where the shape of the drift curve changes from concave upward to concave downward. After the breakdown at 325 derees, the pieces were allowed to cool, and the test chamber temperature was raised to 350 degrees. At this level both specimens drifted rapidly to breakdown in about three minutes.

EFFECT OF GLAZE

All data which have been discussed up to this point have been obtained on unglazed specimens. In order to determine the effect of glaze on the high temperature performance of porcelain, two cylinders were prepared $1^1/_8$ inches in diameter and $2^1/_4$ inches long. One of these cylinders was glazed with a standard glaze of the type normally applied to high voltage insulators. The other cylinder was left unglazed. Silver electrodes were applied and ceramicly bonded to each end of each cylinder. The silver coatings were carried over the ends and extended down the side walls on each end for a distance of $^5/_{16}$ inch. This was done in order to place the highest voltage gradient along the outer glazed surface. Figure 6 shows the results of a simultaneous test made on these two pieces. The resistance of the two was practically the same throughout the test, and both pieces drifted to failure at 410 degrees Fahrenheit. Note that this curve is plotted with total resistance and not resistivity as ordinates.

SOME PRACTICAL INSULATORS

In addition to prepared test specimens, a number of tests have been made on practical insulators of various types and forms. Figure 7 shows the results of two such tests made on a standard 23-kv heavy duty 2-shell outdoor switch insulator. The insulator was tested with its regular metallic parts cemented in place in the usual manner. A direct correct voltage of 14 kv was applied between cap and pin; this voltage is numerically the same as the rms normal service voltage to ground. Note that in Figure 7 the total resistance to ground is plotted as a function of temperature. In this insulator the conduction path from top to bottom metallic elements passes through three layers of portland cement and two layers of porcelain. It should be noted that the initial measurement made at 80 degrees Fahrenheit is considerably below the curve as drawn and also is lower than the second point at 150 degrees. This peculiar performance is believed to be caused by the variation in resistance of the cement. Portland cement is quite hygroscopic and when slightly moist is a fairly low resistance conductor. As the water is driven off, however, its resistance increases so that at elevated temperatures it adds in some measure to the total resistance of the insulator. The high temperature end of the curve is 2-pronged, representing the results from two separate tests. Note that drift curves became measurable on one curve at 350 degrees and on the other

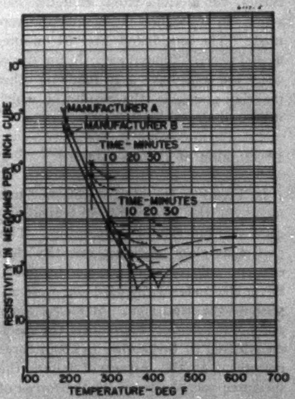

Figure 5. Temperature versus resistivity, comparison of two manufacturers of wet process electrical porcelain

Voltage gradient 20 volts per mil, specimen thickness 0.55 inch

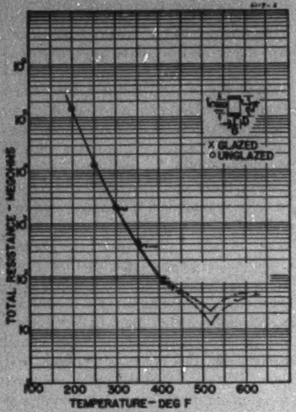

Figure 6. Temperature versus resistance, comparison of glazed and unglazed wet process electrical porcelain

Total voltage applied 5,000 volts direct current

at 400 degrees, reaching stability in each case after about 40 minutes. At 450 degrees the insulators drifted to failure in 20 and 30 minutes, respectively.

A similar insulator was subjected to test at 14 kv rms 60 cycles. This test was made by raising the temperature in 50-degree steps and holding the voltage and temperature constant for one hour at each step. Because of the difficulties of separating conduction from charging currents, no resistance measurements were made. The unit failed after 54 minutes at 400 degrees Fahrenheit. A number of other a-c tests have been made on other porcelain samples, the results of which will be discussed later.

Discussion of Data—Glass

In Figure 8 are plotted the measurements of the resistivity of borosilicate glass as a function of temperature. This particular glass specimen was the glass element only of a suspension insulator. Silver electrodes with guard ring were applied to the head area. At 85 volts per mil and 350 degrees Fahrenheit, the specimen drifted to breakdown in about three minutes. At the instant of failure the specimen was shattered by the forces of differential expansion.

In Figure 9 the variation of resistivity of the same specimen is shown as a function of voltage gradient.

Similarly, Figures 10 and 11 show the performance of a specimen made of a lead glass. This specimen drifted toward failure at 40 volts per mil and 600 degrees Fahrenheit after about three minutes, but upon cooling recovered and was not permanently injured. At 650 degrees and 20 volts per mil, the same specimen showed a definite downward drift in resistivity because of internal heating but stabilized after about ten minutes. At 650 degrees and 30 volts per mil, the piece drifted to failure in about eight minutes. In this case power was not removed quickly enough, and a permanent failure occurred taking the form of a fused path through the glass.

Discussion of Data—Steatite

The results of tests on two steatite specimens are shown in Figure 12. Note that for two voltage gradients, 20 and 40 volts per mil, the curves virtually coincide. Both specimens initially drifted toward failure after 15 and 17 minutes, respectively, at 700 degrees Fahrenheit and 40 volts per mil. The pieces were cooled back to test chamber temperature and recovered. At 750 degrees, both specimens failed at 20 volts per mil after nine and ten minutes, respectively.

There are not sufficient data to plot the variation in resistivity of this steatite composition as a function of voltage for various temperature levels. In Figure 13 such data are plotted for one temperature, only.

Discussion of Data—Zircon

Several zircon compositions were investigated, and it was found that the high temperature electrical characteristics of the several compositions varied radically. The particular zircon composition included in this paper is the one having the best characteristics of the group tried. Figure 14 gives the variation in resistance as a function of temperature with a voltage gradient of 40 volts per mil. Figure 15 shows the variation in resistance as a function of voltage gradient for various temperture.

There are two characteristics of this material which are worthy of particular note.

1. It is observed that in certain ranges increasing voltage gradient results in slight increases in the measured resistance. Such a condition was not observed on any of the other materials, nor even on the other zircon specimens having slightly different composition.
2. It was observed that, at any fixed temperature above about 400 degrees Fahrenheit when a constant voltage was applied, an initial current was measured which diminished with time, finally stabilizing at a value lower than the initial. In other words the measured resistance increased with time. This is exactly opposite to the performance of the other materials. At the lower temperatures, around 500 degrees, this upward drift of resistance reached stability in about one minute with a rise in resistance of about three to four per cent. At higher temperatures, about 800 degrees, the upward drift of resistance reached stability in several minutes with rise in resistance of about 30 per cent. At 800 degrees and at higher voltage gradients, 60 volts per mil, stability was reached only after 20 minutes with a rise of more than 100 per cent of the initial value.

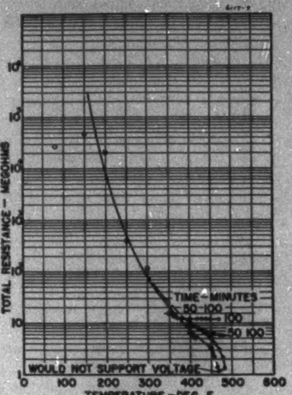

Figure 7. Temperature versus resistance of 23-kv NEMA standard outdoor switch and bus insulator

Total voltage applied 14 kv direct current

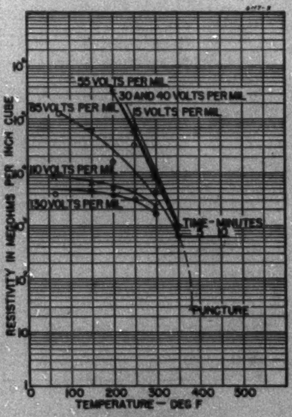

Figure 8. Temperature versus resistivity, borosilicate glass

Thickness 0.36 inch

This characteristic continued up to 900 degrees, the upper limit of the test, and no breakdowns were obtained.

These two unusual characteristics of this material are believed to be caused by dielectric polarization. The curves shown in Figures 14 and 15 are plotted to show the resistance at each level for the polarized or stable condition.

Theoretical Considerations

From an engineering standpoint, it is most important that one be able to determine the temperature and voltage at which a particular insulator may be operated safely without failure. To accomplish this with some reasonable knowledge of the margin of safety, the end point conditions should be predictable. Therefore, some theoretical discussion of the mechanism of breakdown is in order.

From the data it seems reasonably conclusive that, at least at elevated temperatures (250 degrees Fahrenheit and above), breakdown is a thermal phenomenon. When voltage is applied to a dielectric, a leakage or conduction current flows through the material and the I^2R loss results in internal heating above the ambient temperature. This increase in internal temperature results in a reduction of resistivity with consequent increase in internal heating, if a constant applied voltage is maintained. As the internal temperature rises above the ambient, heat is dissipated from the stressed volume. In most practical applications thermal conduction is the principal means of heat dissipation from the stressed volume. The power dissipated as heat from any point in the stressed volume is therefore a function of the internal temperature rise. If the power dissipated as heat increases more rapidly as the temperature rises than does the power input, the time rate of temperature rise will diminish and a condition of stability will be obtained. If the reverse condition occurs, the temperature will rise with increasing rapidity until the dielectric fuses or is is cracked by differential expansion stresses.

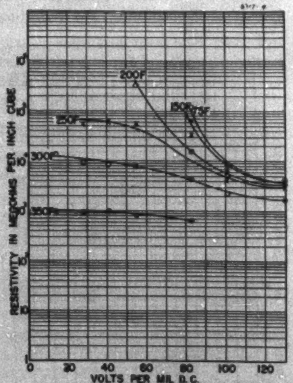

Figure 9. Voltage gradient versus resistivity, borosilicate glass

Thickness 0.36 inch

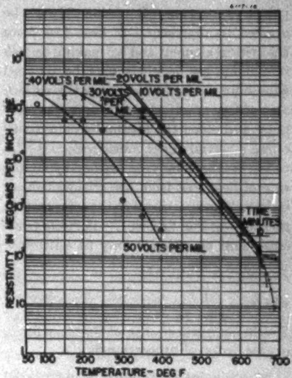

Figure 10. Temperature versus resistivity, lead glass specimen

Thickness 0.50 inch

Consider the elemental cube (Figure 16) in the dielectric body. Let the volume of this cube be ΔV and the edge dimensions be ΔX, ΔY, and ΔZ. Let the direction of voltage gradient be parallel to the Y axis through the elemental cube. If the voltage gradient through the cube in volts per inch is E and if the resulting current density is I amperes per square inch, then P_{T1}, the total power input to the cube, is

$$P_{T1}=(E\Delta Y)(I\Delta X\Delta Z)=EI\Delta V$$

If the resistivity of the material is r ohms per inch cube,

$$P_{T1}=\frac{E^2}{r}\Delta V$$

The power input per unit volume is

$$P_1=\frac{P_{T1}}{\Delta V}=\frac{E^2}{r} \tag{1}$$

If the total power output from the cube in the form of heat is P_{T2}, then

$$P_{T2}=\int_s K\nabla T\cdot \mathbf{n}\,d\sigma$$

where

T = the temperature rise of the cube above the ambient temperature
∇T = the temperature gradient outward from the cube
$d\sigma$ = any element of surface of the cube
$\mathbf{n}$ = a unit vector normal to that element of surface
K = the thermal conductivity of the material

The power output per unit volume is

$$P_2=\frac{P_{T2}}{\Delta V}$$

$$=\frac{K}{\Delta V}\int_s \nabla T\cdot \mathbf{n}\,d\sigma$$

Taking the limit as ΔV approaches zero

$$P_2=K\nabla^2 T \tag{2}$$

The term $K\nabla^2 T$ cannot be evaluated exactly from theoretical considerations alone. However, this term merely represents the power output in the form of heat from a small element in the stressed volume. In any practical insulator, as soon as a voltage is applied some element within the stressed volume begins to heat slightly faster than its immediate surroundings. This may be because the element is at the center of the volume where heat loss is the slowest, because of variations in voltage gradient, or because of variations in the homogeneity of the material. As soon as the element in question begins to rise in temperature above that of the surrounding stressed volume, its resistivity diminishes and consequently its rate of heating is increased further. A "hot spot", therefore, is established quickly, and it is this hot spot which is of

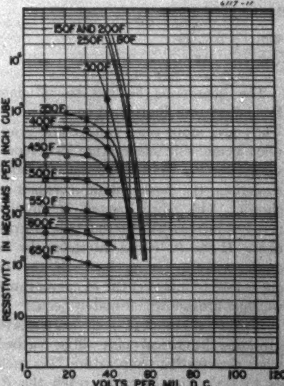

Figure 11. Voltage gradient versus resistivity, lead glass

Specimen thickness 0.50 inch

primary concern. Equation 2 may be applied to represent the heat loss from the critical element and by making certain reasonable assumptions may be reduced to a more useful form. Since the surface area of any practical insulator is relatively large and since the thermal conductivity of the material under consideration is relatively small, it may be assumed that the rate of heat loss from the critical volume is directly proportional to the temperature rise above the ambient. For these assumptions equation 2 may be written

$$P_2 = K\alpha T \tag{3}$$

where

T = the temperature rise of the element
α = the factor of proportionality

At any point of stability P_2 must equal P_1.

When voltage is first applied, $T=0$, and consequently $P_2=0$. As T increases, P_1 and P_2 both increase. If, with increasing temperature T, P_1 continuously increases faster than P_2, stability can never be reached, and the temperature will continue to rise until failure occurs. Conversely, if P_2 continuously increases faster than P_1, a condition of equilibrium will ultimately be reached where $P_2=P_1$.

The rate of increase of P_1 with respect to temperature T is

$$\frac{\partial P_1}{\partial T} = -\frac{E^2}{r^2}\frac{\partial r}{\partial T}$$

Since, for the materials under consideration, $\partial r/\partial T$ is negative, this shows a positive value for the product $-\dfrac{E^2}{r^2}\dfrac{\partial r}{\partial T}$.

The rate of increase of P_2 with respect to temperature is

$$\frac{\partial P_2}{\partial T} = K\frac{\partial(\nabla^2 T)}{\partial T} \cong K\alpha \tag{4}$$

The conditions leading to ultimate breakdown are illustrated in Figure 17. The power input when voltage first is applied at ambient temperature T_0 is represented by P_1. As the internal temperature rises, the power input increases as shown. The power output is initially zero but increases as internal temperature rises. If the P_2 curve intersects the P_1 curve, stability is reached. If the P_2 curve fails to intersect the P_1 curve, runaway and breakdown ultimately occur.

If the power input (P_1) curve is concave upwards, the critical point will be reached when the starting temperature T_0 is such as to make the P_2 curve reach the P_1 curve at a point of tangency. This occurs at some critical temperature T_c and with power input P_1'', as illustrated in Figure 17.

At this point

$$\frac{\partial P_1}{\partial T} = \frac{\partial P_2}{\partial T}$$

or

$$-\frac{E^2}{r^2}\frac{\partial r}{\partial T} = K\alpha \tag{5}$$

The initial ambient temperature T necessary ultimately to reach the critical condition is lower than T_c by an amount equal to P_1'' divided by $K\alpha$.

If the power input curve P_1 is concave downward, the P_2 curve always will intersect it at some point and with some theoretical finite temperature rise above the ambient. Whether or not failure occurs then depends upon whether the temperature at this point of intersection is within the range which the dielectric can stand without fusing, cracking, or charring. It is perfectly possible that the power input curve may be in some zones concave upward and in other zones concave downward with a flex point as illustrated by the broken line in Figure 17.

Equation 5, therefore, may be applied only when the second derivative of power input with respect to temperature is positive at the critical point.

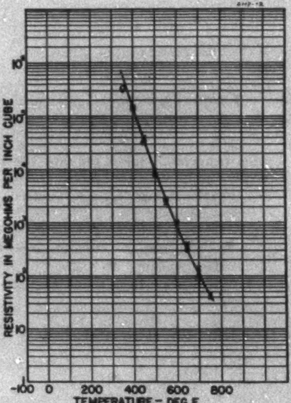

Figure 12. Temperature versus resistivity, wet process steatite

Voltage gradient 20 and 40 volts per mil, specimen thickness 0.188 inch and 0.203 inch

$$\frac{\partial^2 P_1}{\partial T^2} = -\frac{E^2}{r^2}\frac{\partial^2 r}{\partial T^2} + \frac{2E^2}{r^3}\left(\frac{\partial r}{\partial T}\right)^2$$

This term is always positive when $\dfrac{\partial^2 r}{\partial T^2}$ is negative. If $\dfrac{\partial^2 r}{\partial T^2}$ is positive, the term is positive only when

$$\frac{E^2}{r^2}\frac{\partial^2 r}{\partial T^2} < \frac{2E^2}{r^3}\left(\frac{\partial r}{\partial T}\right)^2$$

$$\frac{\partial^2 r}{\partial T^2} < \frac{2}{r}\left(\frac{\partial r}{\partial T}\right)^2 \tag{6}$$

The values of r and of the first and second partial derivatives can be determined physically from the temperature–resistivity curve of any material. For porcelain, the steatite composition, and the two glasses equation 6 is satisfied within the range investigated. For zircon the second derivative of r with respect to T is negative up to about 350 degrees Fahrenheit. Above 350 degrees equation 6 is satisfied up to about 700 degrees. Above 700 degrees, the power input curve is concave downward, and stability always can be reached theoretically at some temperature rise. In this respect above 700 degrees zircon will behave more like a very high resistance conductor than an insulator.

Hence equation 5 can be applied to porcelain, the two glasses, and the steatite composition within the range for which data exist. It also can be applied to the zircon up to 700 degrees Fahrenheit. The values of thermal conductivity K for these materials are given approximately in Table I.

For most of these materials, the thermal conductivity increases somewhat with increasing temperature. By neglecting

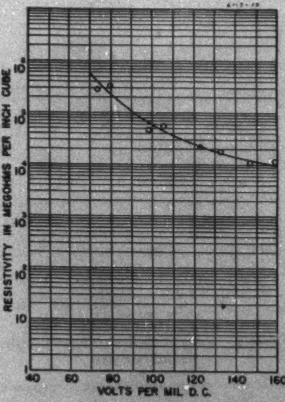

Figure 13 (right). Voltage gradient versus resistivity, wet process steatite

Temperature 84 degrees Fahrenheit, specimen thickness 0.188 inch and 0.203 inch

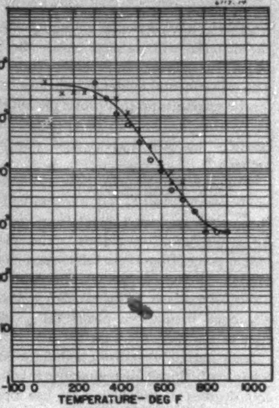

Figure 14. Temperature versus resistivity, zircon

40 volts per mil, 2-square-inch electrode 0.489 inch thick and 1-square-inch electrode 0.196 inch thick

this change using the values given in Table I throughout, conservative results will be obtained.

In order to evaluate equation 5 numerically, the value of the factor α must be determined. The points of ultimate failure with direct voltage for all porcelain test specimens are listed in Table II. The theoretical values for $-\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ are shown for each case. It again should be noted that the data in Table II were obtained by applying constant voltage and raising the ambient temperature in steps. In many cases failure probably would have occurred at some lower temperature if smaller steps had been used. For this reason those tests have been discounted on which failure occurred in less than three minutes. For the remaining tests, the average value of the term $-\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ is 0.067 with a range from 0.0189 to 0.124. The last column in Table II shows a word symbol indicating whether or not the drift curve to failure passed through a flex point. Those tests marked "flex" showed a definite flex point in the drift curve. For those points marked "no", the drift curve continued with increasing acceleration to failure with no flex point. For cases where no flex point occurred in the drift curve, the starting temperature must have been at some point above T_c. For those tests where a flex point appeared, the starting temperature must have been between T_0 and T_c. The values of $-\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ for those tests which passed through a flex point range from 0.0189 to 0.0925, with an average value of 0.0398. For those tests which did not show a flex point on the drift curve, the value of $-\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ ranged from a high of 0.124 to a low of 0.023, with an average of 0.0725. By inspection it appears that the value of the function $-\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ should be about 0.06 at the flex point. If 0.06 is taken as the approximate numerical value of the function at the flex point, then this may be substituted in equation 5, and the factor α may be evaluated as approximately three for porcelain. This evaluation is not very precise, but it should be noted that an error of even three to one in the value of α produces an error of only about 50 degrees Fahrenheit in the theoretical end point temperature.

For porcelain the resistivity varies but little with voltage gradient for temperature above 250 degrees Fahrenheit. A reasonably accurate general solution, therefore, can be obtained for temperatures above 250 degrees. For temperatures below 250 degrees, some confusion exists probably because of the variation of resistivity with voltage gradient, and sufficient data are not available for an accurate analysis. As a matter of interest, equation 5 has been evaluated for porcelain below 250 degrees using the two extremes of the resistivity curve shown in Figure 4.

Table I. Thermal Conductivity

Material	Watts Per Sq In. Per In. Per Deg F
Porcelain	0.0203
Steatite	0.0285
Borosilicate glass	0.0182
Lead glass	0.0182
Zircon	0.031

This yields the theoretical internal temperature T_c at which runaway actually starts for any particular voltage gradient. In Figure 18 the values of T_c have been plotted against voltage gradient. The starting ambient temperature T_0, from which the critical internal temperature T_c ultimately may be reached, also has been plotted in Figure 18 taking

$$T_c - T_0 = \frac{\text{watts per cubic inch input at } T_c}{K\alpha}$$

Any point to the left of the T_0 curve should be safe for continuous operation. Points to the right of the T_0 curve should result in failure if given sufficient time.

In Table III the actual breakdown points for several practical insulators are shown. These values also have been plotted on Figure 18.

For the materials other than porcelain, there are not sufficient data to make a determination of α by the method used

Table II. Breakdown Points of Porcelain Test Specimens, Direct Voltage

Electrode Area, Sq In.	Specimen Thickness, In.	Volts Per Mil D-C	Temperature at Beginning of Failure, Deg F	Resistivity at Beginning of Failure, Megohms Per In. Cube	Time Required for Failure, Min	$\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ Watts Per Cu In. Per Deg F	Characteristic of Drift Curve
1	0.131	382	95	Approximately 200,000	5	0.021	No
1	0.122	410	150	Approximately 50,000	7	0.094	No
1	0.075	440	200	Approximately 58,000	Less than 0.02	0.087	No
1	0.275	136	250	5,000	28.5	0.0925	Flex
1	0.258	194	250	Approximately 3,000	2.5	0.287	No
1	0.605	40	300	495	15	0.077	No
1	0.609	40	300	870	15	0.0442	No
30	0.81	39.5	300	1,630	12	0.023	No
1	0.172	87	300	2,150	7.5	0.084	No
1	0.172	87	300	2,150	7.5+	0.084	No
1	0.797	40	300	892	16	0.043	Flex
2	0.535	20.6	325	516	19	0.0189	Flex
2	0.528	20.6	325	348	19	0.0282	Flex
0.785	0.97	30.8	350	193	3+	0.108	No
0.785	0.98	30.8	350	183	3+	0.114	No
1	0.791	20.2	350	405	44	0.0222	Flex
0.44	0.282	35.5	350	285	4	0.0975	No
0.44	0.282	35.5	350	285	4+	0.0975	No
1	0.172	43.5	350	335	6	0.124	No
1	0.172	43.5	350	335	6+	0.124	No
1	0.81	9.87	400	60	62	0.034	Flex
0.785	0.97	10.25	400	66	15.5	0.0336	No
0.785	0.98	10.25	400	57	15.5	0.0387	No
0.785	0.975	20.4	400	47	2.5	0.186	No
0.785	0.980	20.4	400	47	2.5+	0.186	No

Table III. Breakdown Points of Practical Porcelain Insulator Shapes, Alternating and Continuous Voltage

	Continuous Voltage					Alternating Voltage
	Cylindrical Smooth Walled Standoff Insulator	Cylindrical Corrugated Standoff Insulator	Standard 23-Kv Heavy Duty Outdoor Switch and Bus Insulator	Tube $2\frac{3}{4}$ by $1\frac{3}{4}$ by 20 In. Long, Smooth Walled	3-Unit $5\frac{3}{4}$ by 10 In. Suspension Insulator	Standard 23-Kv Heavy Duty Outdoor Switch and Bus Insulator
Total porcelain thickness, in.	1.62	5.12	1.25	20	2.06	1.25
Total voltage applied, volts	15,000	40,000	14,000	46,000	46,000	14,000 rms
Average volts per mil	9.25	7.8	11.2	2.3	22.3	11.2 rms
Temperature at beginning of failure, deg F	410	400	405	550	350	400
Total resistance at beginning of failure, megohms	79	82	3.26–5.2	24.8	82.2	Approximately 2.7
Approximate average resistivity at beginning of failure, megohms per in. cube	48.7	50.4	81–128	4.36	461	Approximately 66
Time required for failure, min	21	45	17–35	49	112	54
$\frac{E^2}{r^2}\frac{\partial r}{\partial T}$ watts per cu in. per deg F	0.036	0.025	0.0254–0.0162	0.0205	0.0237	0.032
Characteristics of drift curve	Flex	Flex	No	Flex	Flex	—

for porcelain. It is difficult to see why the value of α should differ materially from that for porcelain. If we assume that α is the same for the steatite, glass, and zircon compositions as for porcelain, then theoretical breakdown curves can be plotted for these materials. Such curves for the four other materials are plotted in Figure 18 within the range of the existing data. For borosilicate glass no data area vailable above 350 degrees Fahrenheit, and for temperatures below 250 degrees the resistivity varies greatly with voltage as indicated in Figures 8 and 9. The curve for borosilicate glass, therefore, can be generalized from this data over only a short range. In Figure 18 the borosilicate glass curve is shown in broken lines so that it may be distinguished from the porcelain curve.

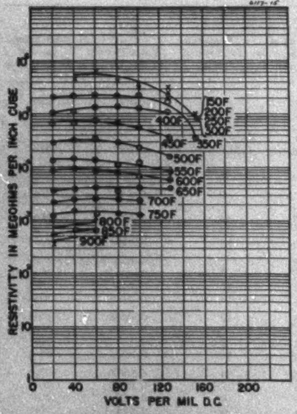

Figure 15. Voltage gradient versus resistivity, zircon

1-square-inch electrode, specimen thickness 0.196 inch

Examination of Figures 10 and 11 will show that the lead glass curve can be generalized from existing data only within the temperature range from 350 to 650 degrees Fahrenheit and for voltage below 40 volts per mil.

No data are available on the steatite composition to show the variation of resistivity with voltage at the higher temperatures. There is every indication, however, that, like porcelain, the resistivity of steatite becomes essentially independent of voltage at temperatures above 250 degrees Fahrenheit. The plot for steatite in Figure 18 is based on this assumption.

For the zircon composition, it appears that above about 140 volts per mil resistivity will vary considerably with voltage. Also, above 700 degrees stability, theoretically, always can be reached. The generalized zircon plot is therefore quite short but does indicate adequately the relative position of this material. Above 700 degrees there is some question as to whether at any particular voltage gradient the material will stabilize before the internal temperature rise becomes excessive. The experimental evidence indicates that for d-c gradients up to 40 volts per mil the zircon composition can be operated successfully at 900 degrees and possibly higher.

As a practical check on these theoretical predictions, Tables II, III, IV, and V give the actual measured breakdown points for a number of specimens. The individual points from these tables have been plotted on the curves in Figure 18 for direct comparison. As would be expected, all actual breakdown points fall to the right of the theoretical curve.

Some actual breakdown points for porcelain are included for temperatures below 250 degrees Fahrenheit. It has been pointed out that at these lower temperatures the generalized curves are of doubtful value. It is interesting to note, however, that the actual low temperature breakdown points are at least of the proper order relative to the theoretical curves. Some breakdown points obtained with alternating voltages also are included in Tables III and IV, and are plotted on the curve in Figure 18. In each case these have been listed and plotted in rms rather than crest volts, since the heating effect is a function of the former. These seem to fit the curves equally as well as do the d-c points, but the amount of a-c data available are relatively small. No actual resistance measurements were made with alternating current and the resistivities listed in Table IV are taken from Figure 4.

Despite the fact that the temperature–resistivity curves were obtained from test specimens with relatively uniform voltage gradients, the theoretical break-

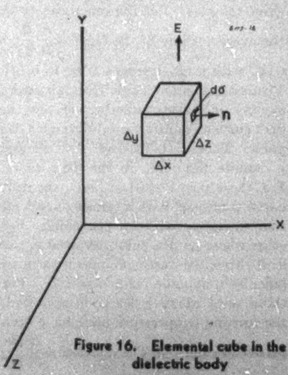

Figure 16. Elemental cube in the dielectric body

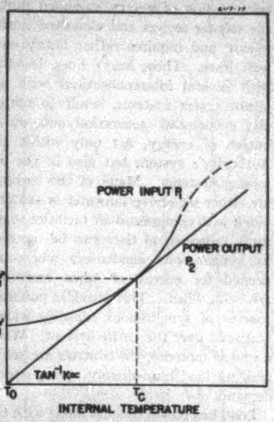

Figure 17. Power input and power output conditions leading to ultimate breakdown

down curve for porcelain applies with reasonable accuracy to practical insulator shapes at the higher temperatures. This is shown by the actual breakdown tests on practical insulators as listed in Table III.

It seems probable that at temperatures above 250 degrees the reduction in resistivity of the porcelain results in more nearly uniform voltage gradients in the practical insulators.

No actual breakdowns were obtained on the zircon composition even though temperatures were carried up to 900 degrees at 40 volts per mil direct current. The tendency of the zircon to increase in resistivity with time at constant ambient temperature and voltage is believed to be caused by dielectric polarization. At any constant temperature and voltage, the more mobile negative ions (or possibly electrons) drift away from the negative electrode, leaving a super abundance of the less mobile positive ions in this area. This results in an apparent increase in the resistivity of the material. When such polarization occurs to a marked degree, it would be expected that the breakdown temperature with alternating current would be lower than with direct current.

Conclusions

The electrical resistivity of all of the ceramic insulating materials tested diminishes not only with increasing temperature but also, at the lower temperatures, with increasing voltage gradient. The rate of change of resistivity with voltage appears to diminish with increasing temperature so that for porcelain above 250 degrees the resistivity is nearly independent of voltage.

From the shape and level of the temperature resistivity curves, it is possible to predict the temperature at which an insulator will fail under any particular voltage gradient. While these predictions are based primarily on high temperature measurements, it appears as though the thermal effect may be a major factor in insulator breakdowns even at low temperatures. The existing data do not permit definite determination at low temperature, however.

Table IV. Breakdown Points of Porcelain Test Specimens With Alternating Voltage

Electrode Area, Sq In.	Specimen Thickness, In.	RMS Volts Per Mil	Temperature at Beginning of Failure, Deg F	Calculated Resistivity at Beginning of Failure, Megohms Per In. Cube	Time Required for Failure, Min	$\frac{E^2}{r^2}\frac{\partial r}{\partial T}$. Watts Per Cu In. Per Deg F
1	0.113	531	−70	600,000	0.25	0.0118
1	0.125	448	−70	1,200,000	5.5	0.0042
1	0.137	438	71	50,000	1.5	0.077
1	0.150	347	77	100,000	18	0.024
1	0.121	346	150	80,000	2.25	0.030
1	0.117	342	250	5,000	0.116	0.469

Table V. Breakdown Points of Glass and Steatite Test Specimens

Material	Electrode Area, Sq In.	Specimen Thickness, In.	Volts Per Mil D-C	Temperature at Beginning of Failure, Deg F	Resistivity at Beginning of Failure, Megohms Per In. Cube	Time Required for Failure, Min	$\frac{E^2}{r^2}\frac{\partial r}{\partial T}$, Watts Per Cu In. Per Deg F	Characteristic of Drift Curve
Borosilicate glass	0.6	0.36	83.3	350	625	3	0.6	No
Lead glass	0.442	0.50	40	600	272	14	0.100	No
Lead glass	0.442	0.50	30	650	110	7.25	0.22	No
Steatite	0.785	0.188	21.3	750	43.9	13	0.188	Flex
Steatite	0.785	0.203	19.7	750	42.5	13	0.166	Flex
Steatite	0.785	0.188	42.5	700	104.5	1.5	0.285	No
Steatite	0.785	0.203	39.4	700	101.3	3	0.252	No

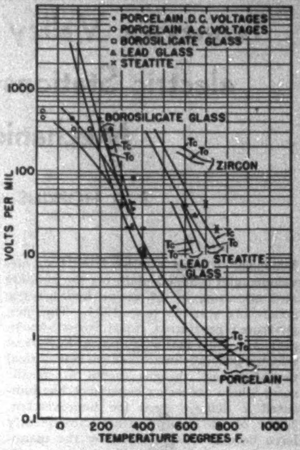

Figure 18. Temperature versus voltage gradient for breakdown

The breakdown curves shown in Figure 18 are believed to be applicable for 60-cycle alternating as well as direct voltage for all of the materials except zircon. For zircon the a-c breakdown temperatures may be somewhat lower.

There is also some question as to whether these curves are applicable to high frequency alternating voltages. At high frequencies other forms of dielectric loss may become of important magnitude, so that failure may occur at lower temperature levels. This has not been investigated.

There seems to be no appreciable difference in the high temperature electrical performance of wet process electrical porcelain as made by several well-known United States manufacturers.

The presence of glaze on porcelain does not seem to affect the high temperature performance.

The high temperature performance of glasses, steatite, and zircons may be expected to vary widely with different compositions. The one zircon composition which thus far appears to be best can be operated at 900 degrees Fahrenheit at 40 volts per mil direct current without failure.

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Tennessee Valley Authority Hydroelectric Stations—Electrical and Mechanical Design

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Synopsis: Design policies have been developed and established during the building and operating of 14 major hydroelectric generating stations since 1934. Turbines, governors, generators, transformers, switchgear, and other major equipment have been selected generally in accordance with normal manufacturers' designs except in specific cases where operating experience has indicated a definite need for improvement. Many of the improvements subsequently have become standardized by the manufacturers. Switching, bussing, control, relaying, wiring, piping, auxiliary equipment, and other station features have been developed by the Tennessee Valley Authority, using manufacturers' standard components as far as possible. The bases for selection of equipment and for development of station designs undoubtedly are applicable to other hydroelectric projects.

A REPORT was made in December 1941[1] on the electrical design of the Tennessee Valley Authority's new hydroelectric generating stations after six of the new stations were in operation. Today with 14 new stations operating, this supplemental report brings the record up to date, adds the mechanical design features, and outlines the fundamental electrical and mechanical design practices which have controlled the new generating station designs.

The Tennessee River system drains an area of 40,910 square miles, lying in the states of Tennessee, Virginia, North Carolina, Georgia, Alabama, Mississippi, and Kentucky. From the highest headwater in the mountains of North Carolina to the lowest tailwater in the western corner of Kentucky there is a total fall of 3,188 feet in 714 river miles. Mean annual rainfall on the watershed averages 52 inches and run-off averages 44 per cent. The unregulated stream flow at Kentucky Dam, the dam nearest the mouth of the Tennessee River, ranged from a minimum of 4,500 cubic feet per second to a probable flood of 960,000 cubic feet per second.

The Tennessee Valley Authority now owns and operates 26 hydroelectric plants with 89 units aggregating 1,777,580 kw, and 12 fuel plants with 37 units aggregating 430,400 kw. By agreement with the Aluminum Company of America, the authority dispatches the water and the power at eight Alcoa dams with a total installed capacity of 311,120 kw of hydroelectric power, and by agreement with Memphis Power and Light Company, it leases 20,000 kw of steam power, thus bringing the total installed capacity now controlled by the Authority to 2,539,100 kw of which 82.3 per cent is hydroelectric.

The principal transmission lines, between generating stations and large load centers (Figure 1), consist of about 2,000 circuit miles of 154-kv lines and 1,000 circuit miles of 110-kv lines. Although the generation and the load are scattered fairly well over the entire area, the hydroelectric generation in the eastern portion of the system is mostly of a storage nature, while that in the central and western portion is generally run-of-river. This condition causes a considerable transmission of energy eastward during the storage season and westward during release and requires rather heavy east-west lines. These heavy lines, together with several interconnections with adjacent power systems, result in unusually economical generation and distribution of energy, not only within the Authority's system but also in the adjoining systems. Many of the turbines are either set above tailwater or are provided with compressed air facilities to depress tailwater so they can be motored as synchronous condensers when not needed for energy or when reservoirs are being filled. This provides potential sources of synchronous capacity widely scattered over the entire system. When a unit is motored, the controls are set to pick up load immediately upon sudden demand.

Load has grown steadily along with the development of generating facilities and much of the time has grown faster than generation could be provided. During the fiscal year ended June 30, 1945, the Authority-controlled plants generated 11,935,000,000 net kilowatt-hours of energy, of which 85.4 per cent was from the hydroelectric station. The maximum hourly demand was 1,933,000 kilowatts.

Generating Stations

The 14 new hydroelectric stations and the six new units at existing Wilson Dam are listed with their principal characteristics in Table I. With their variations in flow, head, storage, ratings, topography, and transmission connections these developments represent three general types of development and several plant arrangements.

Enclosed conventional generator rooms with indoor cranes (Figures 2 and 3) are provided at eight of the 15 plants At Wheeler, where the tailwater is especially low, the generators are outdoors in weatherproof housings, and an outdoor gantry crane serves the main units after removing the housings and serves the erection bays through removable hatch covers. At the remaining six plants an open-deck design (Figure 4) is used with an outdoor gantry crane serving the generating units and erection bays through removable hatch covers.[2]

Transformers are located alongside the

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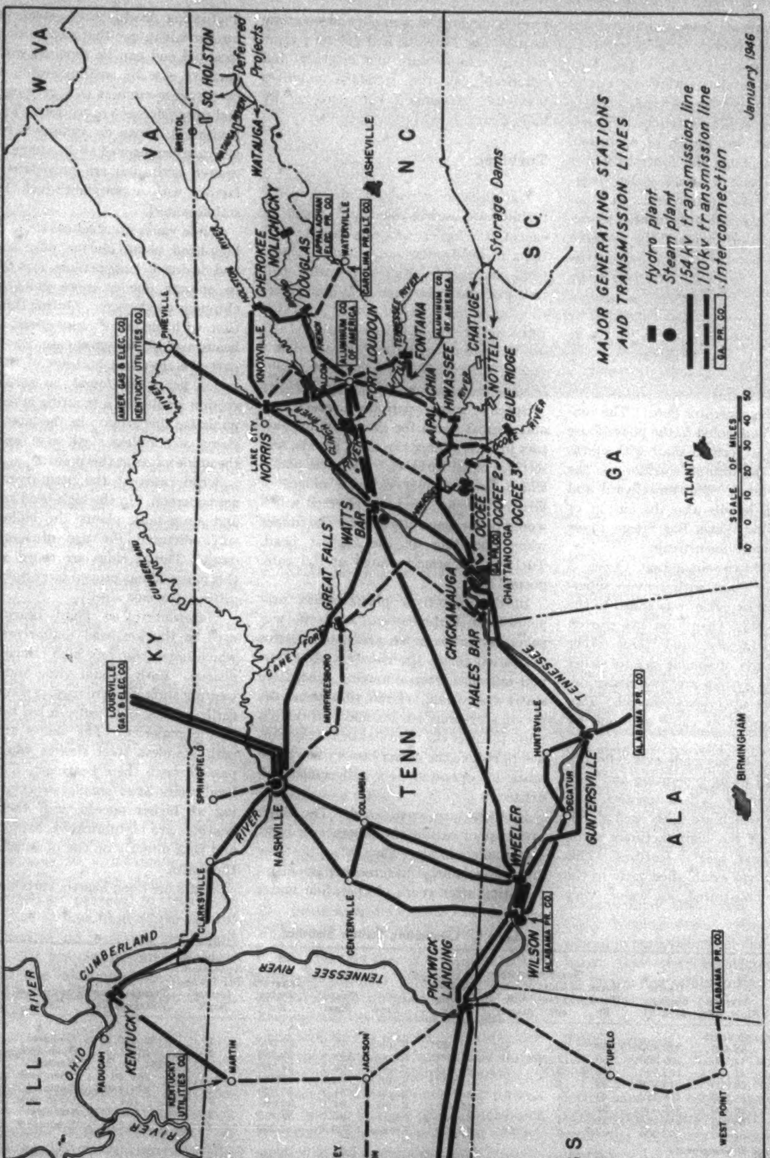
MAJOR GENERATING STATIONS AND TRANSMISSION LINES
Hydro plant
Steam plant
154kv transmission line
110kv transmission line
Interconnection
GA. PR. CO.
January 1946
SCALE OF MILES
10 0 10 20 30 40 50
W VA
VA
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ALA
MISS
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KY
ILL
MO
ARK
SO. HOLSTON
Deferred Projects
BRISTOL
WATAUGA
ASHEVILLE
NOLICHUCKY
CHEROKEE
DOUGLAS
WATERVILLE
APPALACHIAN ELEC. PR. CO.
CAROLINA PR. & LT. CO.
Storage Dams
ALUMINUM CO OF AMERICA
FORT LOUDOUN
FONTANA
CHATUGE
NOTTELY
HIWASSEE
APALACHIA
BLUE RIDGE
OCOEE 1
OCOEE 2
OCOEE 3
KNOXVILLE
PINEVILLE
AMER. GAS & ELEC. CO.
KENTUCKY UTILITIES CO.
LAKE CITY
NORRIS
WATTS BAR
ATLANTA
CHATTANOOGA
CHICKAMAUGA
HALES BAR
GREAT FALLS
CANEY FORK
MURFREESBORO
LOUISVILLE GAS & ELEC. CO.
SPRINGFIELD
NASHVILLE
COLUMBIA
HUNTSVILLE
DECATUR
GUNTERSVILLE
ALABAMA PR. CO.
BIRMINGHAM
WHEELER
CLARKSVILLE
CENTERVILLE
CUMBERLAND
RIVER
TENNESSEE
WILSON
PICKWICK LANDING
PADUCAH
OHIO
KENTUCKY
MARTIN
KENTUCKY UTILITIES CO.
JACKSON
TUPELO
WEST POINT
MISSISSIPPI
RIPLEY
COVINGTON
MEMPHIS
LEABCO
ARK. PR. & LT. CO.
MISS. PR. & LT. CO.

Figure 1. Transmission map, January 1946

generator room at three plants and adjacent to the switching structures at 12 plants. The control room is preferably located between the powerhouse equipment and the switchyard equipment to minimize the lengths of control circuits and to reduce the radius of operation. At 11 of the plants the control room is located in the end or corner of the powerhouse nearest the switchyard. At Apalachia the control room is located alongside the generator room and the governor cabinets are located within the control room, thereby reducing manpower requirements.

At Watts Bar an unusual topography presented a steep rocky bluff at the shore end of the powerhouse (Figure 5). The transformers, the switchyard, and a separate control house are all located 126 feet above the generator floor. The control house is connected to the powerhouse by means of a vertical shaft with elevators and a horizontal corridor at the powerhouse level. This switchyard and control room handle also the output of the 240,000-kw Watts Bar Steam Plant located 3,500 feet downstream.

The 30,000-kva single-unit Ocoee 3 Hydroelectric Plant with remote supervisory control by cable[3] was added to the system in 1943. Operation has proved entirely satisfactory, and three other generating units of similar size for other plants on the system are now under design also for supervisory control. Two will be controlled by cable and one by carrier.

The Authority has made every effort to specify for all of its generating and auxiliary equipment, characteristics and construction which are fairly well standardized by all the manufacturers so as not to demand special features. This policy which was established early in the program demonstrated its value very forcibly during the war when at one time as many as 20 major hydroelectric units were in manufacture and erection, and practically all the standard practices previously developed were successfully maintained.

Turbines

Kaplan-type movable-blade propeller turbines are used in the main-river plants except at Wheeler, where the small variations in headwater and tailwater elevations made a fixed-blade propeller turbine more desirable, and at Wilson where the relatively higher head dictated Francis turbines. Francis turbines are used at all the tributary storage plants and at the tunnel and surge-tank plants because of their high heads.

Specific speed, peripheral coefficient, and critical sigma for cavitation are factors in determining the sizes, speeds, and settings of all turbines.[4] In some plants which operate over extreme ranges of head, a specific speed was selected which would be mechanically safe for the runner when operating at the highest head. Turbines are usually rated at full gate, normal head.

In the main-river plants where propeller-type turbines are used, it was necessary in order to prevent excessive cavitation to set the runners below normal tailwater, thus obtaining a negative static draft head. Draft tube water-depressing systems are installed on certain units to lower the tailwater below the runner to reduce the power losses when these units are operated as synchronous condensers.

Runners are cast steel. The high first cost of cast-steel runners over cast-iron runners is more than offset by their greater resistance to cracking or disintegrating after years of operation under fluctuating loads. Cast steel has the further advantage that cavitation can be retarded and can be repaired with comparative ease by welding, while welding on cast-iron runners is a very expensive and unsatisfactory operation. To retard cavitation, areas on runners which experience has proved to be subject to excessive cavitation are prewelded at the factory with a one-eighth-inch layer of stainless steel.

Guide vanes are steel castings. In the high-head plants during off-peak hours and during storage periods, it is desirable to operate one or more units as synchronous condensers. During this operation, with the guide vanes closed and full headwater in the scroll case, the leakage past the tops and bottoms of the guide vanes has been found to cause some erosion similar to a scouring action. To minimize this erosion in the later plants, strips of stainless steel were welded to the guide vanes at this point.

Scroll cases at the main-river plants are concrete. In the high-head tributary and surge-tank plants the higher pressures dictated the use of steel scroll cases. Throat rings are rolled steel, as this material has proved more resistant to pitting than cast steel.

Water-lubricated main bearings are used in the low-head main-river plants which have relatively high tailwater conditions. Both lignum vitae and rubber wearing surfaces were used in some of the earlier plants, but Insurock has been used in the majority. The shafts are fitted with stainless steel sleeves where they pass through these bearings. The high-head units have smaller wheels operating at higher speeds, and their main bearings are oil-lubricated, babbitt-lined, and bear directly on the polished steel of the shafts.

Shafts are open-hearth, carbon or alloy

Table I. Generating Station Statistics

Station	Average Flow, CFS	Useful Storage, Acre-Ft	Rated Head, Ft	Number of Units at Present	Number of Units Ultimately	Turbine Rating, Hp	Type	Speed, Rpm	Generator Rating, Kw	Number of Generators Per Transformer Bank	Transmission Voltages, Kv	Powerhouse Type	Switchyard Type
Main-River Developments													
Chickamauga	38,410	329,400	36	3	4	36,000	Kaplan	75.0	27,000	1	154, 110, 44	Enclosed	Low
Fort Loudoun	14,300	109,300	65	2	4	44,000	Kaplan	105.8	32,000	2	154, 110	Open-deck	Low
Guntersville	41,400	162,900	36	3	4	34,000	Kaplan	69.2	24,300	1	154, 110, 44	Enclosed	Low
Kentucky	63,620	4,010,800	48	4	5	44,000	Kaplan	78.3	32,000	1, 2	154, 66	Open-deck	Low
Pickwick	55,000	418,400	43	4	6	48,000	Kaplan	81.8	36,000	1	154, 110	Enclosed	High
Watts Bar	26,400	337,600	52	5	5	42,000	Kaplan	94.7	30,000	1	154	Open-deck	Low
Wheeler	50,000	348,900	48	4	8	45,000	Fixed	85.7	32,400	2	154	Outdoor	High
Wilson*	52,000	52,500	92	6	10	35,000	Francis	100.0	25,200	2	154, 110	Enclosed	High
Tributary Storage Developments													
Cherokee	4,650	1,473,100	100	2	4	41,500	Francis	94.7	30,000	2	154, 66	Open-deck	Low
Douglas	6,700	1,419,700	100	2	4	41,500	Francis	94.7	30,000	2	154, 110	Open-deck	Low
Fontana	3,970	1,154,300	330	2	3	91,500	Francis	150.0	67,500	1	154	Enclosed	Low
Hiwassee	2,150	364,700	190	1	2	80,000	Francis	120.0	57,600	1	154, 44	Open-deck	Low
Norris	4,600	2,281,000	165	2	2	66,000	Francis	112.5	50,400	1	154, 110	Enclosed	High
Tunnel and Surge-Tank Developments													
Apalachia	2,240	37,500	360	2	2	53,000	Francis	225.0	37,500	1	154	Enclosed	High
Ocoee 3	1,270	9.400	280	1	1	33,500	Francis	200.0	27,000	1	154, 66	Enclosed	High

* New units in existing plant.

steel forgings, heat-treated, and machine-finished. They have renewable sleeves where they pass through the water-lubricated guide bearings and the stuffing boxes in the head covers. They are all hollow-bored to allow for internal inspection. The upper ends of the shafts for Kaplan wheels are belled out to receive the blade-tilting servomotors.

Governors

The governing system for each turbine is of the hydraulic relay valve actuator type, with motor-driven governor head, auxiliary control mechanism, restoring mechanism, oil-pressure system, pipe, and, in the case of the Kaplan units, the control valve for runner blade adjustment. The governing mechanism and pumping units are mounted in a cabinet (Figure 6), the lower part of which forms the oil sump and the front face of which is the turbine instrument and control board. Governors for either one or two units are mounted in one single or one double cabinet as the powerhouse arrangement dictates.

The governors are designed for a normal maximum operating oil pressure of 300 pounds per square inch. Each governor is provided with a separate vertical cylindrical oil storage tank, with sufficient compressed air space above the oil to operate the turbine wicket gates through one complete cycle. Oil-pressure systems are interconnected, and sufficient spare pumps are provided to permit operation of all governors in the plant during outage of one of the pumping units.

Gate operating time, as controlled by the governor, is adjusted from a minimum of six seconds to a maximum of ten seconds at the various plants. Speed-level controls are provided to control the speed of the turbine from 85 per cent of rated speed at no load and zero speed droop to 105 per cent of rated speed at full load and maximum speed droop. Speed droop adjustment is from zero to six per cent.

Generators

Generator characteristics and construction were developed at the beginning of the program in consultation with the engineers of the manufacturers on the basis of meeting system requirements and at the same time keeping within the manufacturers' normal designs and practices as far as possible. A few compromises were necessary on both sides, and over the years a few improvements have been developed mutually. Principal features which are practically uniform for nearly all the 44 generators installed to date are:

1. Power factor, 0.9.
2. Short circuit ratio, 1.1.
3. Condenser capacity, 55 per cent.
4. Line-charging capacity, 80 per cent.
5. Noncontinuous damper windings giving reactance ratio, 1.35.
6. Flywheel effect based on 30 per cent speed rise with full-load rejection and gate closure in five seconds resulting in 120 to 180 per cent of normal flywheel effect.
7. Excitation voltage, 250.
8. Ceiling, 325.
9. Response ratio, 1.
10. Direct-connected main and pilot exciters partially telescoped.
11. Thrust bearing below the rotor and usually one guide bearing also below the rotor.
12. Shaft flanged at top to rest on thrust bearing and to support the rotor.

An important shop test required by the authority consists of assembling the generator and turbine shafts and their bearing surfaces and checking the complete assembly in a lathe. It is specified that the running surface of the thrust bearing runner shall be perpendicular to the axis of the generator and turbine guide bearing journals within a tolerance of 0.0005 inch at every point. After erection in the powerhouse, each generating unit is given a mechanical rotation check to ensure that the thrust bearing surface is perpendicular to the axis of the generator and turbine guide bearings within the tolerance allowed in the shop test and that the shaft is plumb to within a tolerance of 0.001 inch in about five feet.

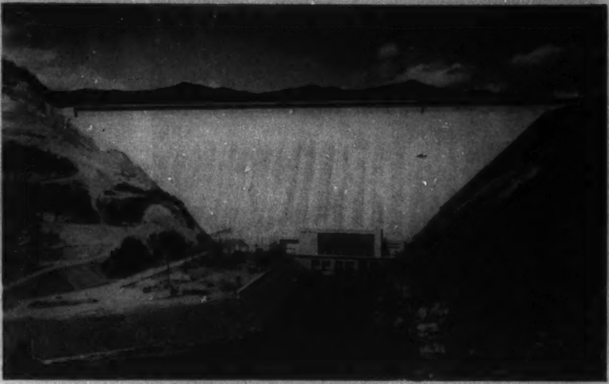

Figure 2. Fontana Dam

Enclosed generator room

Carbon - dioxide fire - extinguishing equipment for the enclosed generators has been improved through the years. Spare cylinders originally were kept on hand to renew an exhausted bank. These cylinders now are racked and piped to form a reserve bank always ready for instant use. A systematic test procedure has been developed wherein a few cylinders are released and the carbon-dioxide concentration is measured over a period of 30 minutes with the generator running. This provides a basis of judging the airtightness of the generator housing and for adjusting the time intervals at which the delayed discharges should be released. The tests also have demonstrated that with a reasonably tight housing fewer cylinders are required in both the initial and the delayed banks than were considered necessary on the earlier units before these tests were developed.

Transformers

Fundamental features originally selected have been carried through to date with minor improvements in keeping with the advances in manufacturing practice.

Three-winding transformers with load-ratio control from the switchboard (Figure 7) have been installed at Ocoee 3 and Kentucky in addition to those previously reported. It has become general practice to provide spare transformers

Figure 4 (above). Kentucky Dam

Open-deck station with double-trolley gantry crane

Figure 3 (left). Chickamauga enclosed generator room

only for important ties and system interconnections where outage would cause loss of load, but not for generation stepup.

Fire protection for transformers has not been installed except at Wheeler, Pickwick, and Wilson, where the transformers are located adjacent to the powerhouse structures. These three systems employ fixed fog nozzles arranged to envelop the transformers. Flow of water is initiated manually by remote control or automatically by thermostats and differential relay operation depending on local conditions.

Main Electrical Connections

Operation to date has indicated the general adequacy of the bussing and switching scheme established at the beginning of the power program. While no fundamental changes have been necessary, a number of minor improvements have become possible as the power system has grown.

At Watts Bar the connection of the four new 60,000-kw steam turbine generators to the hydro bus increased the generation in the Watts Bar area to the point where it no longer was felt necessary to provide separate 154-kv switching for each of the five 30,000-kw generators. Furthermore, space was required in the 154-kv yard for more 154-kv lines than originally had been anticipated. Consequently hydroelectric units 1 and 2, which were already installed, with their individual 3-phase transformers, were paralleled solidly on the high-voltage side and switched as a 60,000-kw unit to the 154-kv bus. Units 4 and 5, installed later, similarly were connected and switched. This provides six 60,000-kw and one 30,000-kw generator circuits on the 154-kv bus with six existing lines and space for two future lines. Synchronizing breakers were installed between hydroelectric generators 1, 2, 4, and 5 and their transformers. The retaining of the individual 3-phase transformers provides flexibility and uniformity, while the group switching to the 154-kv bus conserves breakers and switchyard space.

At Guntersville, where an additional line was needed urgently during war-time restrictions, generators 1 and 2 with their 3-phase transformers, which were already in operation, also were paralleled on the high-voltage side and switched as a 48,600-kw unit to the 154-kv bus. Synchronizing breakers were added between the generators and their transformers. While this conservation of steel structures, high-voltage breakers, and equipment was particularly desirable during the war, the power system had grown to the point where the common switching of the two generating units could be justified as a permanent arrangement.

In a few cases it has been found that system operation does not require as many 154-kv bus-sectionalizing disconnects as originally planned, and fewer disconnects have been provided, particularly in the transfer busses.

The ring-bus arrangement has been used successfully in the initial 154-kv installations at several plants, including Kentucky, Fort Loudoun, Fontana, Douglas, and Cherokee. As pointed out in the previous report, the standard switchyard design provides for a ring-bus arrangement for three or four circuits, and the construction is such that the ring bus is converted readily to a main-and-transfer-bus arrangement when more circuits are added. This conversion has been made in a number of cases.

At Fontana a departure was necessary from the established policy of avoiding local distribution at generator voltage. Since there were no nearby power lines of a voltage lower than 154 kv, and since there was a demand of some 3,000 kw in village load, it was considered expedient to avoid excessive stepdown transformation by taking the village service from the 13.8-kv delta busses of the two stepup transformer banks of generators 1 and 2. A 13.8–12.5-kv transformer bank and a step-type regulator in the village feeder, together with station-type arresters, guard the generator circuits against voltage surges from lightning.

Generator-Voltage Switchgear and Connections

Air blast circuit breakers have been given consideration during recent years, but up to the present time conventional oil circuit breakers have been used exclusively. The conventional top-connected breakers are well adapted structurally to the Authority's standard switchgear designs, and because of the small number of indoor breakers required, never more than two per generator, the oil fire hazard is not considered serious.

Cables from generators to switching cubicles for many of the later installations have been insulated with varnished-cambric tape instead of corona-resisting rubber compound. They are shielded with copper tape. The ducts generally are drained and ventilated and in this case the cables are provided with an outer cover of tape and braid. In any case where moisture may be expected, the cables are covered with Neoprene jackets. Creepage terminals are formed in much the same way as with rubber insulation except that particular attention is given to the excluding of all moisture by finishing the exposed insulation with several coats of Glyptol varnish and by filling the end of the stranded conductor with solder. The cables are designed for operation at a conductor temperature not exceeding 78 degrees centigrade.

Cables from switching cubicles to transformers are in most cases paper-insulated and lead-covered, as previously reported, but two exceptions will be noted. At Watts Bar Dam, where a 126-foot difference in elevation exists between the two ends of the circuits, varnished-cambric insulation with nonmigrating compound is used instead of oil-impregnated paper insulation. At Wilson Dam and at Kentucky Dam, where distances up to 1,475 feet were required and where the circuits were carried through areas where conventional duct lines would have congested available space, Oilostatic cable circuits were installed. The Oilostatic circuit, which consists of paper-insulated cables with the three phases installed in a steel pipe filled with oil, used about 75 per cent of the copper and 50 per cent of the space which would have been required for conventional duct lines.

Bare structural shapes have been used to advantage in three recent installations where rather heavy circuits were required between the switching cubicles and the transformers. At Apalachia, one $2^{1}/_{2}$-inch iron-pipe-size copper tube per phase was used for the 1,927-ampere circuits outdoors. At Fort Loudoun, two $2^{1}/_{2}$- by-$2^{1}/_{2}$- by-3/16-inch copper angles per phase forming a square with vertical ventilation were used for the 3,420-ampere circuits on the ceilings of concrete tunnels with Transite barriers and bottom covers and with forced ventilation at full load. At Fontana two $3^{1}/_{2}$- by-$3^{1}/_{2}$- by-1/4-inch copper angles per phase forming a square with vertical ventilation were used for the 3,610-ampere circuits on the side walls of concrete tunnels with Transite barriers and front covers and with natural ventilation. These three installations represent copper economies of 1,220, 1,950, and 1,100 amperes per square inch, respectively.

Transmission-Voltage Switchgear and Connections

Oil circuit breaker ratings on the 154-kv system have been improved at many of the older stations from 2,500,000 kva, 8-cycle opening, 35-cycle reclosing, to 3,500,000 kva, 5-cycle opening, 20-cycle reclosing, with pneumatic operators; and recently all new breakers have been purchased with the higher ratings. Likewise, the few 1,500,000-kva breakers originally installed at certain earlier stations have been improved to 2,500,000 kva.

Air blast breakers have been given considerable study, but during war restrictions it was necessary to use the conventional oil breakers, and so far they have been able to meet all essential requirements.

Automatic reclosing at 20 cycles without synchronism check now is being used on some of the 154-kv tie lines.

Currents in excess of 600 amperes have developed on certain 154-kv heavy tie lines and busses, requiring the continued use of 1,200-ampere breakers, the reinforcing of certain disconnecting switches, and the changing of some of the steel pipe conductors to copper tubing.

Lightning Protection

The arresters located at the terminals of the stepup transformers now are rated 145 kv for the 154-kv system, 121 kv for the 110-kv system, and 73 kv for the 66-kv system. At Apalachia, the 154-kv switchyard where the lines terminate is approximately 1,000 circuit feet from the arresters at the transformers, and consequently station-type arresters are provided also in the switchyard. These arresters are rated 169 kv, which is considered adequate for the protection of the switchgear.

Figure 5. Watts Bar project

Two elevations

Relaying

Generator protection against faults to ground, particularly near the neutral ends of the stator windings, has been improved in some of the later stations by use of a low-energy short-time overcurrent relay connected as a differential ground relay to compare the residual current in the generator main leads with the current in the generator neutral lead. This relay is now used in place of the ground backup relay with low-current long-time setting previously used.

Most of the 154-kv lines now are provided with distance phase relays and directional overcurrent relays controlled by carrier pilot relays. The directional overcurrent ground backup relays also are used as originally. The recent type GCY and HZM relays are being used on the longer lines.

Annunciation

The operation recorder previously mentioned now has become standard equipment at all new generating stations. It records on a strip chart the important manual and switching operations and all abnormal conditions usually annunciated. One or two recorders, depending on the station requirements, are located on the operator's desk alongside the telephone turret (Figure 8). With the recorder equipment, it is no longer necessary to provide the large array of annunciator windows along the top of the instrument board as installed at a few of the earlier plants. The particular annunciations which require immediate attention of the switchboard operator are arranged in small windows on the recorder housing on the operator's desk, and those requiring immediate attention of the turbine operator are arranged in windows on the governor actuator panels.

Communication

Power line carrier telephone channels now provide direct communication between all generating stations and the dispatcher's office. Most of the stations are also connected to the dispatcher's office by telephone lines which are owned or leased by the authority. These telephone lines carry both voice and carrier channels.

Telemetering by power line carrier recently has been installed for transmitting the load on each tie connection with other utilities from the point of interconnection to the dispatcher's office in Chattanooga. Here the loads are recorded, totalized, and retransmitted to the dispatcher of one of the neighboring utilities or are used to control automatically the impulses of a master system load controller. These impulses are broadcast over the entire Tennessee Valley Authority system by a frequency-shift type carrier equipment and can be received at all of the larger hydroelectric stations for regulating the governors of local generators in response to load changes on a distant tie-line interconnection. The total generation at the major stations also is transmitted to the dispatcher's office by means of the carrier telemetering channels. Equipment is being installed at the two largest generating stations, Wilson and Watts Bar, to enable these two stations to carry a definite proportion of the regulating load when operating in parallel under the control of the master controller. These stations then will share the load in the same manner as the generators within a station share the load when operating under proportionate load control.

Carrier-current equipment now is being located generally in the air-conditioned telephone rooms, although earlier carrier relay sets were located in the switchyard. The inspection and servicing of this type of equipment are facilitated greatly by having the equipment indoors.

Switchboards and Controls

Within the control room the general arrangement of boards, panels, and equipment, as previously described, has become standardized with few exceptions. At Apalachia, a 2-unit particularly compact station, and at Ocoee 3, a 1-unit unattended station, the controls and instruments are mounted on a common vertical board, and a bench is not required.

Switchboards in the more recent stations (Figure 9) instead of being black are finished in generally light colors blending with the finish of the control room, while all equipment has contrasting standard black finish. This combination presents a very pleasing appearance and improves vision. Draw-out type relays with built-in test facilities are becoming standard.

Wiring of the switchboards constantly is becoming simpler and more accessible. In most of the later stations, except the largest ones, terminal blocks have been mounted very successfully on wing panels extending back from main panel intersections, and the control cables are carried through rectangular steel ducts to asbestos trays in the room below. This eliminates the 3-story arrangement, with terminal cabinets on the intermediate floor, which is used at the larger plants. At the plants where the terminal blocks and other devices are located in cabinets on the floor below the switchboard, the wiring was made continuous from the terminals on the switchboard equipment to the terminal blocks and was therefore installed by the Authority. With the terminal blocks on the wing panels, the wiring can be installed either at the factory or in the field. In cases where time has been available to get all wiring requirements into the hands of the manufacturer in time for him to do the wiring, it has been done in the factory; but in many cases of extreme urgency, time has been gained by having the boards shipped unwired and installing the wiring in the field. Control cables recently have been insulated with oil-base compound containing Buna S and have been jacketed with Neoprene.

Automatic oscillographs have proved very useful in analyzing line faults, and in stations having a large number of lines, two 6-element oscillographs have been installed.

Station Auxiliaries

Indoor overhead cranes of capacities up to 300 tons are provided in all stations having conventional enclosed generator rooms for erection, dismantling, and maintenance of the generating units. They also are used for handling transformers in those stations where the transformers can be rolled into the generator room for untanking. Most of these cranes have two trolleys, each with a main hoist and an auxiliary hoist. A lifting beam on the two main hooks is used to handle the heavier loads. Bridge drive is by two motors mounted near the center of the bridge and connected through individual cross-craneway shafts to at least 50 per cent of the wheels on each rail. One drive is provided with a hydraulic-operated service brake and the other with a time-delay thrustor-released spring brake. Trolley drive is by one motor mounted on the trolley connected to one wheel at each side of the trolley and provided with a solenoid-operated

Figure 6. Double governor cabinet at Watts Bar

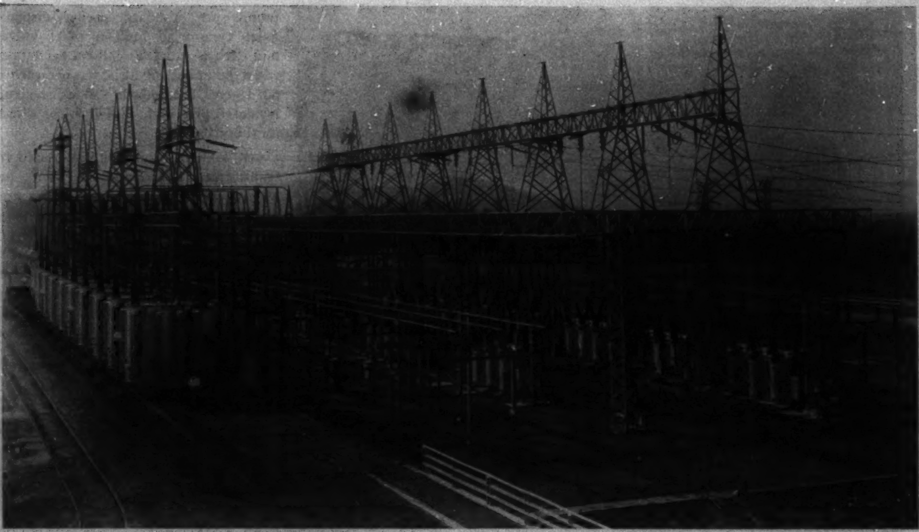

Figure 7. Three-winding transformers and low-type switchyard at Chickamauga

brake. The main and auxiliary hoists have individual motors which are identical and interchangeable and are provided with thrustor-released spring brakes. All motors are 440-volt 3-phase ball or roller bearing crane type. All are wound-rotor type except the trolley drive motors which are squirrel-cage type. Controls of bridge travel, trolley travel, and hoisting are arranged for independent control of these three motions and to permit only one motion at a time. Acceleration is by adjustable definite-time automatic control. The control provides for movements from rest of less than 1/16 inch for the main and auxiliary hoists and less than 1/4 inch for the bridge and trolley travels. The controls are of magnetic type with master switches in the operator's cab. Power supply for the crane is at 440 volts, 3-phase, by steel or copper angles in the runway and sliding shoes on the crane. Power and control connections to the trolleys are by steel angle conductors on the bridge girders and sliding shoes on the trolleys.

Light overhead single-beam cranes, of capacities up to five tons, operated from the floor by cord switches are provided in several of the erection bays and machine shops.

Unloading cranes and derricks are provided at a number of the plants for receiving materials coming in by river.

Outdoor double-trolley gantry cranes (Figure 4) of capacities up to 275 tons are provided over the generator rooms and erection bays at Wheeler and at all stations of the open-deck design for erection, dismantling, and maintenance of the generating units. These cranes also are used for handling transformers in those stations where the transformers can be rolled under the gantry crane and lowered into the erection bay for untanking. These gantries are in most cases provided with overhanging bridges or with boom hoists to permit handling of tail gates. Each of the two trolleys has a main hoist and an auxiliary hoist. Gantry drive is by means of four motors, one at each corner of the gantry and operated through a common control. Trolley drive is by one motor mounted on each trolley and connected to one wheel on each side of the trolley. Each of the four hoists has an individual motor, all four motors being identical. The motors, brakes, and controls are essentially the same as for the indoor overhead cranes. Power supply for the gantry is by flexible cable and motor-driven cable reel from polarized receptacles installed along the runway except at Kentucky, where collector angles are used. Power and control connections to the trolley are by steel angle conductors on the gantry and sliding shoes on the trolleys. All machinery is protected from the weather by a steel housing, and particular attention was given to the streamlining of the crane structures and the housings, since the large gantry is one of the prominent features at each site.

Outdoor single-trolley gantry cranes of capacities up to 100 tons are provided at most of the stations for handling spillway gates, intake gates, stop logs, emergency bulkheads, trash gates, and trashracks. An important design practice at most of the stations is to build the deck at the same elevation across the spillway and intake sections so that the cranes can operate over both these sections. The trolley has a short travel over the two slots of a gate. It is provided with a hoist and with latching attachments for handling the gates. Drive motors, brakes, and controls are essentially the same as for the double-trolley gantry, and the machinery is covered. Power supply to the gantry is normally by means of conductors under deck and collector shoes on the crane. In emergency, power is supplied by flexible cable and motor-driven reel from polarized receptacles installed along the runway. Power and control connections to the trolley are by flexible cable looped between a terminal box on the gantry and one on the trolley.

Spillway gates[8] at Watts Bar, Fort Loudoun, Cherokee, and Douglas are of

Figure 8. Control desk with operation recorders and telephone turret at Chickamauga

Figure 9 (right). Control room with asymmetrical indirect lighting and colored switchboards at Fort Loudoun

the radial type and are raised and lowered by means of two traveling hoists (Figure 5) running on rails on the deck for the full length of the spillway. Each hoist consists essentially of two chain drums driven by a single motor, the whole mounted on a 4-wheeled carriage. The hoist motor is squirrel-cage type, and the control makes use of regeneration while lowering. A separate squirrel-cage motor is provided for driving the carriage. The equipment is enclosed in a streamlined housing. Power supply is by flexible cable and motor-driven reel from polarized receptacles installed along the runway. At Apalachia Dam, where radial spillway gates also were used, individual fixed hoists were installed in order to provide remote control at a later date.

Intake gates at high-head plants are of the roller type and are operated by fixed hoists located in chambers near the tops of the dams. They are controlled by local push buttons, and emergency close buttons are provided at the governor control boards.

Compressed air[4] at 100 pounds per square inch is provided in all stations for tools, generator brakes, various pneumatic controls in the ventilating system, and turbine grease lubrication. The usual practice is to install one stationary 105-cubic-foot-per-minute compressor and one portable 100-cubic-foot-per-minute compressor in each plant. Several plants are provided with additional air compressor capacity to depress the tailwater below the turbine runners while motoring the units and in one plant to prevent the formation of ice on the spillway gates. Black steel uninsulated pipe with welded or screwed fittings is used for all air lines.

One kind of lubricating oil is used both for the governor pressure system and for bearing lubrication. It has a viscosity of 300 Saybolt-second units at 100 degrees Fahrenheit. The oil is purified by a 350-gallon-per-hour centrifuge provided in each plant. Two tanks are provided, one for clean oil and one for dirty oil. When space permits, these tanks are located in the powerhouse adjacent to the oil-purification room. Where no such space is available, the tanks are buried in the fill outside the powerhouse wall. One clean-oil pump and one dirty-oil pump are furnished to handle the oil between the purifier, the tanks, and the generating units. The pumps are standard positive-displacement type with built-in relief valves, and each is rated 30 gallons per minute when pumping oil at a temperature of 50 degrees Fahrenheit. Black steel uninsulated pipe with welded joints is used for the oil lines.

One kind of insulating oil is used for transformers and circuit breakers, but oil once used in a breaker is never used in a transformer. The insulating oil is purified and handled (Figure 10) by a separate system in each plant similar to the lubricating-oil system. The purifier is a 900-gallon-per-hour combination centrifuge and filter press with electric heater, and the pumps are 100 gallons per minute each. The supply and return lines are not connected permanently to the transformers and breakers, but a valve is located in each line near the equipment, and flexible hose is used when filling, draining, or circulating.

Runner hub oil is required for lubricating the runner blade operating mechanisms in the Kaplan turbines. This oil has a viscosity of approximately 1,700 Saybolt-second units at 100 degrees Fahrenheit. Each hub contains about 1,500 gallons of oil. Whenever the runner is dismantled for inspection or repair, it is necessary to remove this oil, and it is salvaged for reuse.

Portable pumping facilities and tanks are provided to handle and transport this oil, but no purification equipment is installed.

Raw water is used in coolers for the main generators, the air compressors, and the air-conditioning equipment. In some plants it is used for fire protection, lubricating the turbine guide bearings, and as makeup for the treated-water

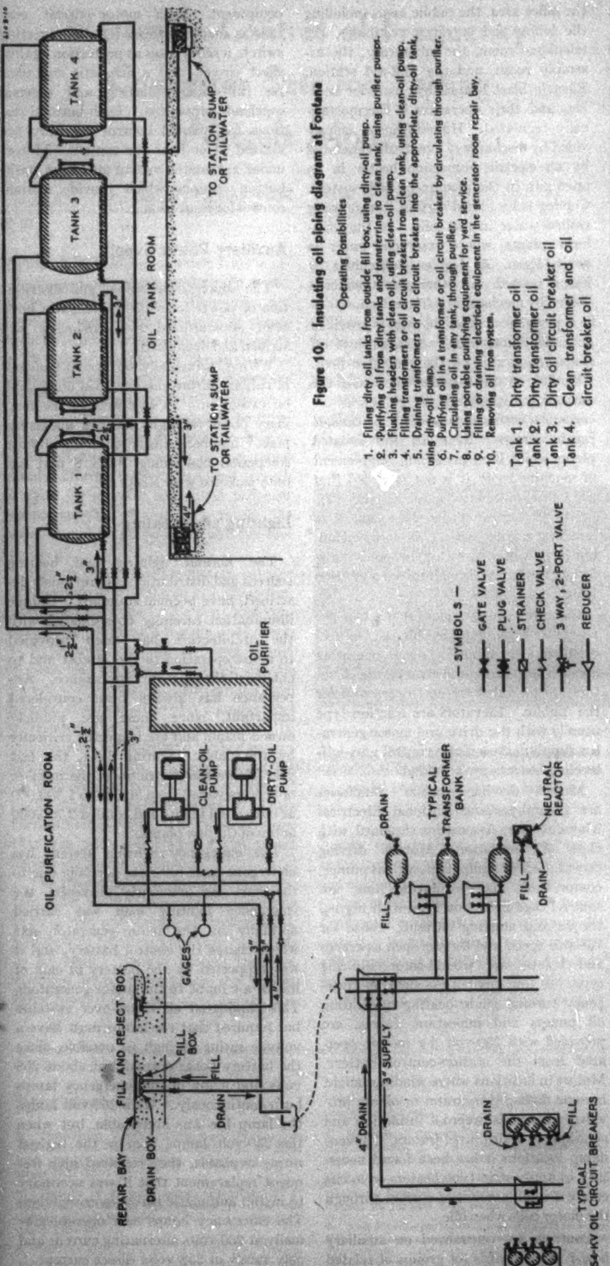
OIL PURIFICATION ROOM
OIL TANK ROOM
TANK 1
TANK 2
TANK 3
TANK 4
TO STATION SUMP OR TAILWATER
TO STATION SUMP OR TAILWATER
OIL PURIFIER
CLEAN-OIL PUMP
DIRTY-OIL PUMP
GAGES
FILL AND REJECT BOX
FILL BOX
FILL
DRAIN
REPAIR BAY
DRAIN BOX
4" DRAIN
3" SUPPLY
DRAIN
FILL
TYPICAL TRANSFORMER BANK
NEUTRAL REACTOR
FILL
DRAIN
DRAIN
FILL
TYPICAL 154-KV OIL CIRCUIT BREAKERS
2½"
3"
4"
— SYMBOLS —
GATE VALVE
PLUG VALVE
STRAINER
CHECK VALVE
3 WAY, 2-PORT VALVE
REDUCER

Figure 10. Insulating oil piping diagram at Fontana

Operating Possibilities

1. Filling dirty oil tanks from outside fill box, using dirty-oil pump.
2. Purifying oil from dirty tanks and transferring to clean tank, using purifier pumps.
3. Flushing headers with clean oil, using clean-oil pump.
4. Filling transformers or oil circuit breakers from clean tank, using clean-oil pump.
5. Draining transformers or oil circuit breakers into the appropriate dirty-oil tank, using dirty-oil pump.
6. Purifying oil in a transformer or oil circuit breaker by circulating through purifier.
7. Circulating oil in any tank, through purifier.
8. Using portable purifying equipment for yard service.
9. Filling or draining electrical equipment in the generator room repair bay.
10. Removing oil from system.

Tank 1. Dirty transformer oil
Tank 2. Dirty transformer oil
Tank 3. Dirty oil circuit breaker oil
Tank 4. Clean transformer and oil circuit breaker oil

system. In the low-head plants, water is taken from the scroll case in each unit and pumped through a multiple basket strainer, having 3/16-inch openings, to a header leading to the station system. The pumps are horizontally split double-suction single-stage volute-type centrifugal units. Specifications require enclosed bronze impellers with wearing rings, bronze shaft sleeves through the stuffing box, water seal piping, and deep stuffing boxes. Experience has proved that this type of pump will give years of service with little maintenance except occasional repacking and lubrication of the bearings. For the high-head plants, a separate intake from each unit penstock is connected to a central manifold from which the individual systems are supplied through pressure-reducing valves as required. An air cushion chamber is installed on the manifold to absorb any water hammer caused by the operation of the system.

Treated water is used in the various hydroelectric plants for all sanitary services and in some plants for fire protection. Wherever possible, treated water is obtained from a nearby community, even at the expense of laying a mile or two of pipe line. Where this is not possible, filter plants are installed. At projects requiring considerable water to supply an operators' village as well as the hydroelectric station, a conventional-type rapid sand gravity filter plant with mixing chamber, coagulation basin, dry chemical feeders, and chlorinators is usually provided. At projects where the demand is light, pressure-filter plants are installed. System pressure often is obtained by use of elevated storage tanks which usually are left standing from the construction activities at the project.

The station sumps in most plants are unwatered by two 300-gallon-per-minute float-controlled deep-well-type vertical-shaft turbine pumps operating in echelon. This type of pump keeps the motor above the wet conditions of an open sump and prevents possibility of flooding out caused by high tailwater conditions. Very little sand or abrasive matter is present in the water, so enclosed or semi-enclosed impellers are specified. The pumps are bronze-mounted and are oil-lubricated automatically by a solenoid valve energized with the motor circuit. In plants where sufficient head is available from the reservoir, water jet eductors are usually installed in lieu of pumps. Since the pressure water for operation of eductors comes from the unit penstocks and since it is possible for the penstocks to be empty, drainage eductors are usually supplemented by one motor-driven sump pump.

Each draft tube is connected to the station sump with a large size line containing a sectionalizing valve. Most plants have two deep-well-type vertical unwatering pumps which range in capacity from 2,500 to 5,000 gallons per minute according to the volume of the draft tube. They are operated manually and can augment the station sump pumps in an emergency. In high-head plants, water jet eductors are usually provided for unwatering.

Fire protection uses raw water at all plants where the minimum lake level provides sufficient pressure. A pressure of 40 pounds per square inch at the hose nozzle has been adopted as the minimum for this service, but a pressure of 60 to 70 pounds per square inch is preferred and is used when obtainable. In plants where the minimum pressure is not obtainable from the lake, fire protection is obtained from the treated-water system or by means of a hydropneumatic system employing a raw-water booster pump. Valves and hydrants with hose racks are provided at strategic points in the powerhouse and in the switchyard. Fog-type hose nozzles which produce a highly atomized spray are provided for use around electrical hazards.

Portable carbon-dioxide extinguishers of 4 to 15 pounds are provided in the powerhouse, and wheeled truck extinguishers of 200-pound capacity are provided in the switchyards. Fixed automatic carbon dioxide systems, with banks of cylinders similar to those provided for the generators, are installed for the protection of oil-treating rooms, oil-storage rooms, and the gas-engine generator rooms.

Heating to 72 degrees Fahrenheit is provided only for those areas where attendance is required. Other spaces are heated only as required for relief of dampness and excessive chill. All heating is by electric resistance heaters except for the air-conditioned areas. Heaters in general are thermostat-controlled fan-type 440-volt 3-phase units with auxiliary on-off-auto-manual switches. The ventilation is provided in several areas of each plant according to the need for human comfort, dissipation of heat from equipment, dissipation of solar radiation, and relief of dampness.

Air conditioning is provided in certain spaces of each powerhouse for the protection of delicate electrical equipment and for human comfort. The general areas served by air conditioning include the control room and its related spaces, the office area, the public areas including the lobby and entrance vestibules, the telephone room, the laboratory, the assembly room, and the first-aid station. Electric blast heaters are used for heating, and their operation is thermostatically regulated. Humidification is provided by discharge of live steam generated by an electric immersion heater in an open pan in the plenum of each system. Cooling is by finned surface coolers using chilled water in the coils. In the high-head plants where reservoir water is available at 55 degrees Fahrenheit or less, it is used. In other plants, water is chilled by mechanical refrigeration.

A permanent machine shop is provided at each plant, usually located just off the erection area on the generator floor. The size of the machine shop and the amount of equipment are chosen in accordance with the proximity of outside repair service. Even in the isolated plants which have maximum complement of machine tools, it is not intended that they shall be able to make all major repairs. A central repair shop capable of handling a great variety of work, including large water-wheel repairs and transformer rewinding, is maintained at Wilson Dam.

Elevators are provided at most of the plants where considerable differences exist in elevation between operating areas. In several projects separate elevators are provided for employees and for the public. Elevators are traction type usually with d-c drive and motor-generator supply. Automatic control and self-leveling features are provided.

Motors driving station auxiliaries are general-purpose National Electrical Manufacturers Association standard, with class A insulation. Motors driving constant-speed equipment, such as pumps, compressors, and ventilating fans, are squirrel-cage induction type with normal torque and starting current; those for variable speed and torque, such as cranes and hoists, are wound-rotor slip-ring type. A few auxiliaries, such as emergency turbine guide bearing lubricating oil pumps and important valves, are provided with 250-volt d-c motors operated from the station-control battery. Motors in locations where windings might become damaged by water or oil are provided with moistureproof insulation and splashproof or dripproof frames. In very damp locations it has been found necessary to apply strip-type heaters or to circulate a small amount of current through the motor coils when idle.

Controls are centralized on auxiliary power switchboards for groups of related equipment. Each motor circuit contains a manually operated disconnecting switch, a set of fuses as protection against short circuit, and a magnetic contactor for full-voltage starting and thermal overload protection. Push-button stations for manual control generally are located at or near the motors. Motors under automatic control also have push-button stations which provide manual control for inspection.

Auxiliary Power Supply

The sources, switching, and distribution of power for station auxiliaries have been standardized as previously described and illustrated.

Wiring, principally on account of loss of rubber during the war, has undergone an evolution. As an outcome, the auxiliary power wiring in most of the recent plants has been insulated with oil-base compound containing Buna S and has been jacketed with Neoprene.

Lighting and Heating

The normal lighting and heating sources and distribution, as previously described, have become standard. Unique illumination carefully co-ordinated with the architecture has been provided to assist operation to the utmost and to enhance the general appearance. Appearance has always been considered important, since these are publicly owned plants and are inspected critically by the public. During 1941, the last year before war restrictions, the number of visitors varied from 26,000 to 1,300,000 at individual plants and totaled 2,700,000 at ten of the new plants.

The emergency lighting system has undergone some revision, partially due to the war. As previously discussed, the emergency lighting load was carried normally by the motor-generator sets which charge the control battery, and it was supported by the battery in case of loss of a-c input to the motor-generators. This eliminated all throw-over switches but required that the lamps must have a voltage rating as high as possible, since the battery floats normally at about 258 volts and many of the emergency lamps burn continuously. With 260-volt lamps the lamp life was acceptable, but when the 250-volt lamps became the highest rating available, they required such frequent replacement that it was necessary to install automatic throw-over switches. The emergency lamps now operate normally at 230 volts alternating current and only rarely at 258 volts direct current.

Electronic Control of Frequency Converter Sets for Testing Aircraft Models

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THE TESTING of airplane scale models in wind tunnels has brought about a demand for variable frequency a-c power for controlling the speed of power model motors used to drive the propellers of the plane models. To obtain aerodynamic data on the plane model which are equivalent to the expected performance of the actual plane, it is necessary to drive the model propellers at speeds greatly in excess of propeller speeds of the actual plane. During tests in the wind tunnel, it must be possible to control propeller speed and to measure accurately the speed as well as power input into the propeller.

Squirrel cage induction motors, capable of speeds up to 80,000 rpm at 1,330 cycles, have been developed. To start and control the speed of such motors, it is necessary to have available a power supply system the frequency of which can be varied and regulated. To permit model motors of various sizes and manufactures to be used, the a-c output voltage of the system must be adjustable. When using a motor over a wide frequency range, the ratio of voltage to frequency (volts per cycle) must be held constant to obtain constant air gap flux. Thus it is necessary that the volts per cycle of the power supply system be regulated.

The frequency and voltage supply system must be fast in response, unvarying in frequency and voltage at the test point, and flexible for maximum variety of tests to be made. Such a system frees the operator to concentrate on the various other quantities to be held constant during the course of the test, such as air velocity, air temperature, and many others.

The aircraft designer is interested in speed and torque input into the propeller. Torque may be measured directly on the model, or it may be deduced from the electric input data by calibrating the model motors. A set of calibration curves for each motor is then taken on a dynamometer or torque stand, correlating input watts from the variable frequency system and output torque at various speeds.

Figure 1. View of United Aircraft wind tunnel in East Hartford, Conn., at which power model testing equipment is installed

Use of variable frequency a-c power systems is not restricted to plane model testing. Research work on propellers, either inside or outside of a wind tunnel, superchargers, gas turbines are but a few examples of applications which require high speed motors with a wide range of speed control.

Scope

It is the object of this paper to describe a frequency converter installation at the United Aircraft Corporation wind tunnel laboratory operated by the Research Department at East Hartford, Conn. Variable frequency power is utilized for model testing in the wind tunnel, for propeller testing inside and outside the tunnel, and for testing work in the thermodynamics laboratory. Electronic regulators are used for the following control purposes:

1. To adjust and regulate frequency.
2. To adjust and regulate volts per cycle.
3. To maintain parallel operation of two converter sets.
4. To maintain current limit during acceleration and deceleration.
5. To limit alternating voltage available at high volts per cycle settings.

Power Conversion Equipment

Figure 3 is a 1-line diagram of the power circuits of the system. Variable frequency power is supplied to the utilization network by two frequency converter sets *A* and *B*, each consisting of:

One 300-kw 24-pole 1,700-rpm wound rotor converter, 440-volt 3-phase 60-cycle primary, 720-volt 3-phase 400-cycle secondary.

One 400-horsepower, 1,700-rpm 250-volt d-c adjustable speed shunt wound motor.

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The authors wish to acknowledge that the system of providing variable frequency by means of induction frequency converter sets with electronic field control was first proposed by C. C. Clymer, industrial engineering division, General Electric Company.

Figure 2. Two 375-horsepower dual rotation model motors mounted on torque stand in calibration room

Two sets are used so that two independent tests can be run when the combined output of the system is not required.

Output frequency of the converters is varied between 65 and 400 cycles by varying the speed of the d-c driving motors between 25 and 1,700 rpm. Variable voltage control of motor speed is employed. Operation at standstill, or 60-cycle output, is prevented by the control in order to avoid burning of the slip rings. The rotor of each converter contains the primary windings, and the stator the secondary windings. The stator windings have two circuits per phase, the total volts per cycle range is 0.5 to 4. With parallel connection of the circuits of each phase, a range of 0.5 to 2 volts per cycle is obtained; with series connection, the range is one to four volts per cycle. The range selection is made by manual knife switches, to suit the rating of model motors under test.

To supply power to and control the converter sets, a 6-unit constant speed motor generator set is used, consisting of:

One 800-horsepower 1,200-rpm 2,300-volt 3-phase 60-cycle synchronous driving motor.

Two 250-kw 250-volt shunt-wound d-c generators to supply variable armature voltage to the d-c motors of sets *A* and *B*.

Two 156-kva 440-volt 3-phase 60-cycle alternators to supply excitation to the frequency converters.

One 20-kw 125/125 volt compound wound d-c exciter to supply a constant voltage control bus.

Synchronous motor and d-c motor fields are energized from the constant voltage d-c exciter bus. The field excitation of the d-c generators is supplied by thyratron rectifiers from the 60-cycle power system. The field strength, which determines the speed and output frequency of the converters, is controlled by electronic regulators. Similarly, the field excitation of the alternators is obtained from thyratron rectifiers with electronic regulators, providing control of the volts per cycle of the converter output. Included in Figure 3 is a block diagram of the principal electronic field control circuits.

Power Distribution System

Output power from the frequency converters is fed through a distribution system to various points of utilization. From each converter power flows through a line switch and a transfer switch to either the wind tunnel or the thermodynamics laboratory. In the wind tunnel laboratory, each converter bus (*A* and *B*) is split into three channels. Thus six channels are available in the tunnel for testing a model having six propellers. One channel off each bus can be connected to the calibration room, and another pair of channels can be routed to a small pilot tunnel for preliminary test work. Knife switches are used to select the channels.

For tests requiring the combined output of both converter sets, the two converter busses can be paralleled through a paralleling contactor. During parallel operation, voltage and frequency of the system is determined by the control of set *A* which acts as the master. Set *B* then automatically matches set *A* and acts as the follower.

Overcurrent Protection

As the impedance of the converters and the distribution system is comparatively high, the short-circuit capacity of the system is low, and contactors are adequate for interrupting short-circuit current. The short-circuit current flowing in the secondary winding of either frequency converter reflects itself in the primary winding in inverse ratio of coil turns. Instantaneous overload relays, connected in the primary side of the converters, trip out line contactors 1*MA* and 1*MB*. This removes excitation from the converters and reduces their secondary voltage to zero.

Excess power output in the secondary circuit of either converter results in a high current in the loop circuit of the corresponding d-c machines. Therefore, instantaneous overload relays in the loop circuits trip d-c line contactors 2*MA* and 2*MB*.

Protection against sustained overloads is obtained by thermal relays with resistance type heaters in each of the six channels. Each relay is connected in the secondary side of a current transformer. Since the heaters carry only a small amount of current, they are wound of fine wire. The effect of frequency within the system working range on the relay tripping current is negligible. These relays protect the system, and, by proper selection of heater sizes, also protect the model motors.

Metering

In order to obtain an accurate record of the power input into the model motors, precision wattmeters are provided which permit measuring power with an accuracy of repetition of 0.10 per cent. Because of the wide range in current and voltage measured, the wide range in frequency, and the high accuracy required, multitap current and potential transformers are used.

As it is not necessary to meter all channels simultaneously, a separate metering channel may be connected to each bus in series with any one of the three channels. For instance, if contactor 1*A* is closed, channel 1 is connected directly to bus *A*. If contactors 1*B* and 1*C* are closed and contactor 1*A* is open, channel 1 is connected through the metering channel. Control is so interlocked that only one channel can be connected to either metering channel at one time.

In addition to the precision wattmeters, switchboard type voltmeters and ammeters are connected to the instrument transformers. At the beginning of a test, the precision instruments are disconnected, and the operator selects the instrument transformer ratio so that the pilot instruments are within proper reading range. Pilot wattmeters, connected to each metering channel through their own sets of instrument transformers, give the tester a direct indication of power.

Precision tachometers, actuated by tachometer generators built into the model motors, are used to read model motor speed during a test. As these instruments are apart from the power system, their details will not be discussed in this paper.[1] Frequency of the system is indicated by switchboard type d-c voltmeters, calibrated in cycles and connected to tachometer generators which are coupled to the converter sets.

Volts per cycle is measured by a volts per cycle meter which can be connected to either converter. This instrument is an ammeter which measures the current flowing through an inductance, the impedance of which is proportional to frequency. Switchboard type ammeters indicate the load on either converter.

Electronic Control

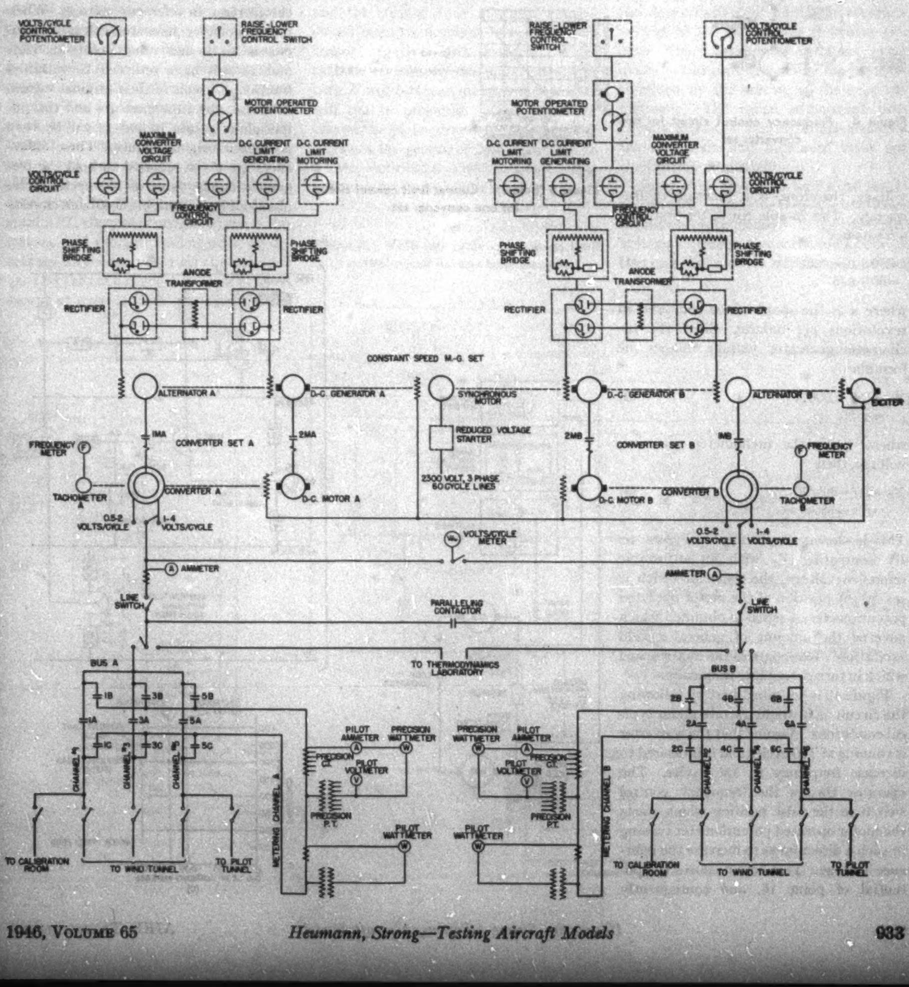
Of first importance in the testing of a model motor is the frequency and voltage applied to the motor. These quantities are adjusted and held constant during a given test by comparing each quantity to a reference voltage and maintaining the difference between the quantity and its reference as small as possible. For example, in the case of frequency control, a d-c tachometer generator directly connected to the converter produces a direct voltage which is a function of frequency. This voltage is matched against a reference voltage in the electronic circuit to hold frequency constant as long as the reference voltage is unchanged.

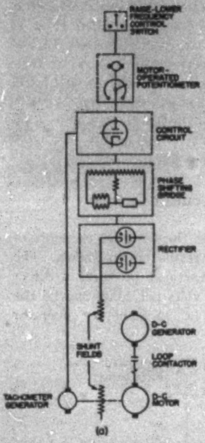
Each d-c generator field and each alternator field is excited by a pair of thyratron tubes in a biphase half-wave circuit in which the thyratron firing angle is controlled by resistance variable inductance bridge circuits. The inductance of the bridge is in the form of a saturable reactor whose d-c winding is supplied from a regulated d-c source and controlled by a high vacuum tube. This is shown in Figures 4, 5, 6, 7, and 8, which are simplified diagrams of the various control circuits. Each of the quantities to be adjusted or held constant is controlled by a closed loop regulating system.

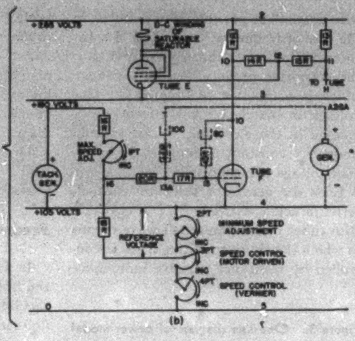
Frequency Control Circuit

In Figure 4a the essential elements of the frequency regulators are shown. The tachometer generator produces a direct voltage of 50 volts per 1,000 rpm of the converter. The tachometer generator

Figure 3. One-line diagram of power model supply system


VOLTS/CYCLE CONTROL POTENTIOMETER
RAISE-LOWER FREQUENCY CONTROL SWITCH
MOTOR OPERATED POTENTIOMETER
MAXIMUM CONVERTER VOLTAGE CIRCUIT
D-C CURRENT LIMIT GENERATING
D-C CURRENT LIMIT MOTORING
VOLTS/CYCLE CONTROL CIRCUIT
FREQUENCY CONTROL CIRCUIT
PHASE SHIFTING BRIDGE
ANODE TRANSFORMER
RECTIFIER
CONSTANT SPEED M.-G. SET
ALTERNATOR A
D-C. GENERATOR A
SYNCHRONOUS MOTOR
D-C. GENERATOR B
ALTERNATOR B
EXCITER
1MA
2MA
CONVERTER SET A
REDUCED VOLTAGE STARTER
2MB
CONVERTER SET B
1MB
FREQUENCY METER
TACHOMETER A
CONVERTER A
D-C. MOTOR A
2300 VOLT, 3 PHASE 60 CYCLE LINES
D-C. MOTOR B
CONVERTER B
TACHOMETER B
0.5-2 VOLTS/CYCLE
1-4 VOLTS/CYCLE
VOLTS/CYCLE METER
AMMETER
LINE SWITCH
PARALLELING CONTACTOR
BUS A
TO THERMODYNAMICS LABORATORY
BUS B
1B
3B
5B
1A
3A
5A
1C
3C
5C
2B
4B
6B
2A
4A
6A
2C
4C
6C
CHANNEL #1
CHANNEL #3
CHANNEL #5
CHANNEL #2
CHANNEL #4
CHANNEL #6
METERING CHANNEL A
METERING CHANNEL B
PILOT AMMETER
PRECISION WATTMETER
PRECISION C.T.
PILOT VOLTMETER
PRECISION P.T.
PILOT WATTMETER
TO CALIBRATION ROOM
TO WIND TUNNEL
TO PILOT TUNNEL


RAISE-LOWER FREQUENCY CONTROL SWITCH
MOTOR-OPERATED POTENTIOMETER
CONTROL CIRCUIT
PHASE SHIFTING BRIDGE
RECTIFIER
D-C GENERATOR
SHUNT FIELDS
LOOP CONTACTOR
TACHOMETER GENERATOR
D-C MOTOR
(a)


+285 VOLTS
D-C WINDING OF SATURABLE REACTOR
TUBE E
2
12R
13R
10
14R
12
15R
11
TO TUBE H
3
+180 VOLTS
13A
16R
MAX. SPEED ADJ.
1PT
INC
10C
9C
20R
17R
16
13
TUBE F
TACH. GEN.
GEN.
4
+105 VOLTS
18R
REFERENCE VOLTAGE
2PT
MINIMUM SPEED ADJUSTMENT
3PT
SPEED CONTROL (MOTOR DRIVEN)
4PT
SPEED CONTROL (VERNIER)
0
5
(b)

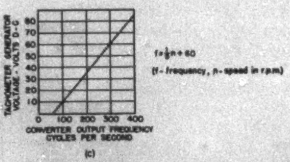


Figure 4. Frequency control circuit for one converter set

voltage, therefore, is a function of frequency. The 24-pole converter produces a frequency

$$\begin{aligned} f &= 60+24n/120 \\ &= 60+n/5 \end{aligned} \qquad (1)$$

where n is the speed of the converter in revolutions per minute. Since the tachometer generator voltage follows the formula

$$\begin{aligned} V_T &= (50/1,000)n \\ &= n/20 \end{aligned} \qquad (2)$$

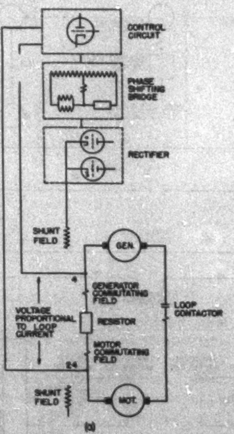
where V_T is the tachometer generator voltage, then

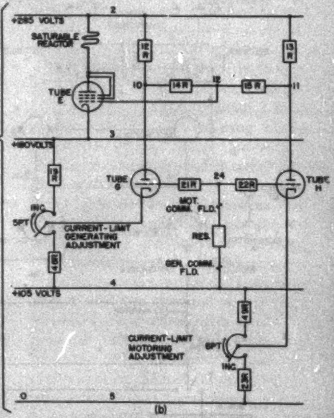
$$\begin{aligned} V_T &= 5(f-60)/20 \\ &= f/4-15 \end{aligned} \qquad (3)$$

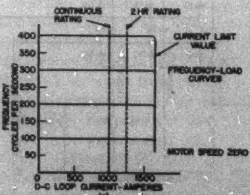
This is shown graphically in Figure 4c. By comparing V_T with an adjustable reference voltage, the value of which is set by the position of the motor operated potentiometer, a signal is obtained which governs the amount of generator field excitation. This controls the motor speed which in turn determines frequency.

Figure 4b is described best by following the circuit as it would operate under typical conditions. Assume that the converter is running at 100 cycles and it is desired to increase frequency to 150 cycles. The operator throws the frequency control switch to the raise position which starts the motor operated potentiometer turning in such a direction as to increase the reference voltage. This action causes the potential of point 16, and consequently point 13 to become more negative with respect to the cathode of tube *F*, which decreases the plate current of this tube. Neglecting for the moment the resistor capacitor circuits in phantom and assuming the potential of point 11 to be held constant at approximately 40 volts positive with respect to bus 4, decreasing the plate current of tube *F* raises the potential of point 10, and likewise the potential of point 12 with respect to bus 3 and cathode of tube *E*. As the plate current of tube *E* increases, the reactor becomes more saturated and the thyratron firing angle is advanced. As the generator field excitation is thus increased, the converter set accelerates to a new speed. This speed is determined by the amount the tachometer generator voltage must increase to offset the increase in reference voltage. When the tachometer generator voltage has increased to its new value, points 16, 13*A*, and 13 will have returned to within a fraction of a volt of their original values. Likewise, the tube currents and the potentials of points 10 and 12 will be very near their original values. These differences are those required to hold the frequency at the value of 150 cycles. The phantom lines represent feedback circuits

Figure 5 (below). Current limit control circuit for one converter set


CONTROL CIRCUIT
PHASE SHIFTING BRIDGE
RECTIFIER
SHUNT FIELD
GEN.
4
GENERATOR COMMUTATING FIELD
LOOP CONTACTOR
VOLTAGE PROPORTIONAL TO LOOP CURRENT
RESISTOR
MOTOR COMMUTATING FIELD
24
SHUNT FIELD
MOT.
(a)


+285 VOLTS
2
SATURABLE REACTOR
12R
13R
TUBE E
10
14R
12
15R
11
3
+180 VOLTS
19R
TUBE G
21R
24
22R
TUBE H
MOT. COMM. FLD.
INC
5PT
CURRENT-LIMIT GENERATING ADJUSTMENT
RES.
GEN. COMM. FLD.
4
+105 VOLTS
CURRENT-LIMIT MOTORING ADJUSTMENT
6PT
INC.
0
5
(b)

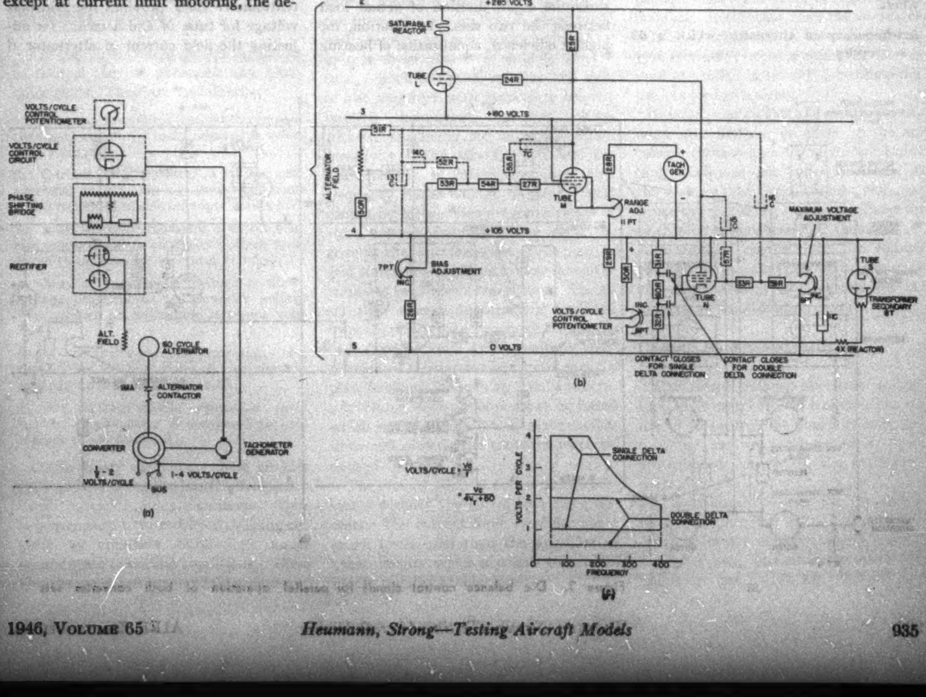


which assure stable operation at all times and maximum speed of response with minimum overshoot under transient changes of load or frequency.

Current Limit Control Circuit

In any system consisting of a d-c motor energized from a d-c generator very high current may flow in the loop circuit between the generator and motor armatures. Although instantaneous overcurrent protection is provided to shut down the motor in the event of high overcurrents it is sometimes desirable to prevent the existence of currents above a chosen value. Figure 5a presents the method of regulating current at the current limit value. An *IR* drop proportional to loop current is compared with two adjustable reference voltages in the control circuit. When power flow is from generator to motor, point 24 is positive with respect to point 4, and tube *G* (Figure 5b) becomes the amplifier for the current signal. When point 24 rises in potential, as the loop current increases, to a point approximately one volt negative with respect to the cathode of tube *G*, whose potential is determined by the current limit adjustment 5*PT*, the plate current of tube *G* increases, and the potential of point 10 decreases. Since the potential of point 11 is fixed by the tube *H* which is turned full on except at current limit motoring, the decrease in potential of point 10, and consequently of point 12 with respect to bus 3, will decrease the plate current of tube *E* thus reducing the generator field excitation as described under the foregoing heading. If the generating current limit action is brought about by rapid motor acceleration, the effect of reducing the generator field by the current limit control circuit is to retard the rate of acceleration of the motor.

If the motor is running at a given speed and the frequency reference voltage is suddenly reduced, the generator field voltage will be decreased, and the motor acting as a generator will supply power to the generator acting as a motor. Under this condition the potential of point 24 becomes negative with respect to point 4, and the current limit motoring control circuit begins to function. As point 24 becomes negative with respect to the cathode of tube *H*, the plate current of tube *H* will decrease, and points 11 and 12 will rise in potential. This action increases the plate current of tube *E*, which increases the generator field excitation to a value sufficient to hold the loop current at the preset current limit value.


VOLTS/CYCLE CONTROL POTENTIOMETER
VOLTS/CYCLE CONTROL CIRCUIT
PHASE SHIFTING BRIDGE
RECTIFIER
ALT. FIELD
60 CYCLE ALTERNATOR
1MA
ALTERNATOR CONTACTOR
CONVERTER
TACHOMETER GENERATOR
½ - 2 VOLTS/CYCLE
1-4 VOLTS/CYCLE
BUS
E
+285 VOLTS
SATURABLE REACTOR
25R
TUBE L
24R
3
+180 VOLTS
51R
ALTERNATOR FIELD
50R
14C
13C
52R
53R
54R
56R
7C
27R
TUBE M
28R
TACH GEN
+
-
RANGE ADJ.
11 PT
16C
15C
MAXIMUM VOLTAGE ADJUSTMENT
4
+105 VOLTS
7PT
INC
BIAS ADJUSTMENT
26R
29R
30R
31R
60R
32R
57R
33R
58R
INC.
8PT
VOLTS/CYCLE CONTROL POTENTIOMETER
TUBE N
9PT
INC
11C
TUBE S
TRANSFORMER SECONDARY 8T
5
0 VOLTS
4X (REACTOR)
CONTACT CLOSES FOR SINGLE DELTA CONNECTION
CONTACT CLOSES FOR DOUBLE DELTA CONNECTION
VOLTS/CYCLE = Vc/f
= Vc/(4VT+60)
VOLTS PER CYCLE
1
2
3
4
0
100
200
300
400
FREQUENCY
SINGLE DELTA CONNECTION
DOUBLE DELTA CONNECTION

(a)

(b)

(c)

Figure 6. Volts per cycle and voltage limit control circuit for one converter set

Volts Per Cycle Control Circuit

The volts per cycle ratio is controlled by comparing the converter output voltage to the tachometer generator voltage which is a function of frequency. In this case the tachometer generator voltage is the reference determined by the frequency control setting. Figure 6a shows a 1-line diagram of this circuit, and Figure 6b the control circuit. Since the frequency is proportional to the tachometer generator voltage, plus a constant, some means must be provided to add a constant direct voltage to the tachometer generator voltage. By connecting the grid of tube *M* to a point sufficiently negative with respect to bus 4, the cathode of tube *M* will be negative in the control range. This has the desired effect of adding a constant voltage to the tachometer generator circuit. The constancy of volts per cycle over the complete frequency range is determined by the setting of the bias adjustment. The range adjustment provides a means of varying simultaneously the maximum and minimum volts per cycle ratios obtainable.

Tube *N* in Figure 6b limits the maximum voltage available from the converter regardless of frequency. As the converter voltage increases, the cathode of tube *N* becomes less positive with respect to the grid, until the tube begins to pass current which restrains the alternator field, thus limiting the converter voltage to a predetermined value.

Figure 6c presents the range of volts per cycle obtainable at each frequency. The curved portion indicates the maximum volts per cycle obtainable at the higher frequencies because of limitations imposed by the maximum converter voltage rating of 720 volts. For the single delta connection, maximum volts per cycle is 4.0 up to 180 cycles and $720/f$ above 180 cycles. For the double delta connection, maximum volts per cycle is 2.0 up to 360 cycles and $720/f$ above 360 cycles.

Parallel Operation

Figure 7 shows the method of comparing *IR* drops in the d-c generator motor loop circuits to balance the currents in these circuits during parallel operation of the converters. The d-c *IR* drop is a true measure of converter load, since at a given frequency the division of power between the d-c motor and the alternator, neglecting losses, is fixed in accordance with the formulas:

$$\frac{\text{D-c motor power}}{\text{Converter power}}=\frac{fc-fa}{fc}$$

$$\frac{\text{Alternator power}}{\text{Converter power}}=\frac{fa}{fc}$$

where

fa = frequency of alternator, which is 60 cycles

fc = frequency of converter, which may be 65 to 400 cycles

For example, when the converter is running at 150 cycles and supplying 75 kw to a load, the alternator load is

$$\frac{60}{150}\times 75=30\text{ kw}$$

and the d-c motor load is

$$\frac{150-60}{150}\times 75=45\text{ kw}$$

Thus, since the ratio of d-c motor load to total converter load is fixed, by the converter frequency, the d-c loop current, which is a function of d-c motor load, is a measure of total converter load. When the two converters are running in parallel, the load on converter *B* is made equal to the load on converter *A* by making the d-c loop current of set *B* equal to the d-c loop current of set *A*.

Figure 7b indicates an adjustment included to balance the two loop currents. This adjustment also provides a means of introducing a small direct voltage to compensate for the grid to cathode voltage of tube *F* in the control range.

Power Factor Balance—Parallel Operation

When the two converters are running in parallel, the d-c current balance circuit insures the equal division of actual load between the two sets. In addition, for highest efficiency, equalization of heating and maximum pull out torque, some means must be provided to balance the power factor of the two converters. There are several methods of accomplishing power factor balance. Possible methods are as follows:

1. Balance total load current of one converter against the other.
2. Balance reactive component of load current of one converter against the other.
3. Balance total current of one alternator against the other.
4. Balance the reactive component of load current of one alternator against the other.
5. Balance the d-c field current of one alternator against the other.

During parallel operation with equalized power and power factor, all of the foregoing currents will be balanced, however, the last mentioned proved to be most satisfactory from the standpoint of stability and circuit simplicity. This is shown in Figure 8.

Figure 8b gives the method of measuring field current by using the voltage drop across a resistor in series with the alternator field. The strength of alternator *A* field is determined by the volts per cycle setting, and the control circuit of set *B* acts to amplify the difference between the two alternator field currents and to control alternator *B* field in such a direction as to reduce the difference to zero. Potentiometer 12*PT* provides the grid bias voltage for tube *M* and a means for adjusting the field current of alternator *B*

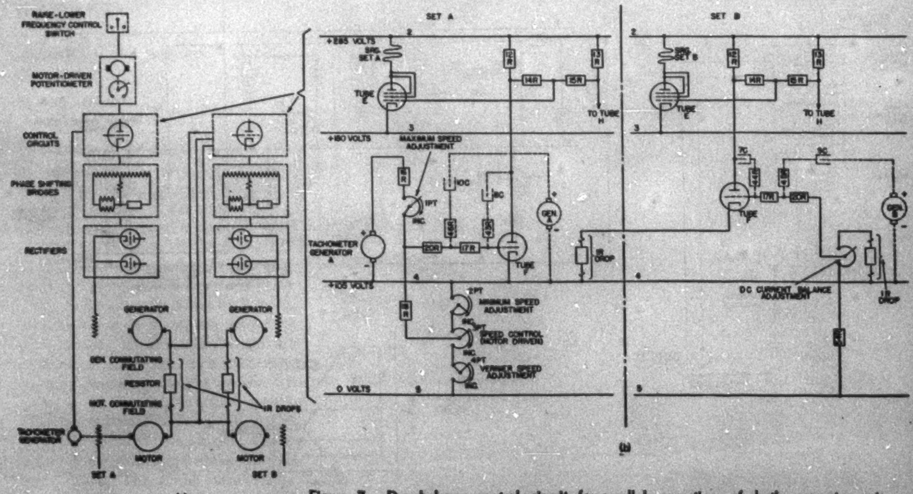
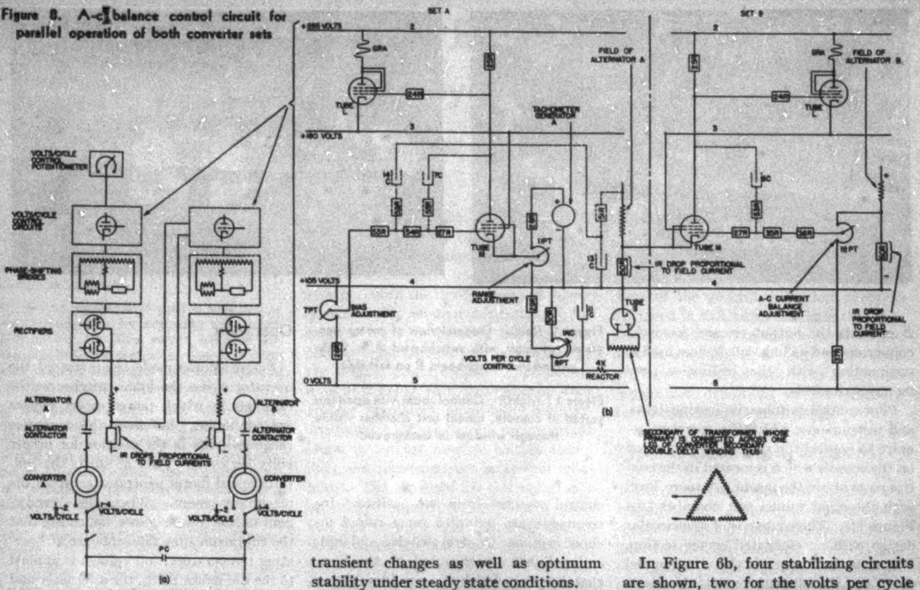

RAISE-LOWER FREQUENCY CONTROL SWITCH
MOTOR-DRIVEN POTENTIOMETER
CONTROL CIRCUITS
PHASE SHIFTING BRIDGES
RECTIFIERS
GENERATOR
GENERATOR
GEN. COMMUTATING FIELD
RESISTOR
MOT. COMMUTATING FIELD
IR DROPS
TACHOMETER GENERATOR
MOTOR
MOTOR
SET A
SET B
(a)
SET A
SET B
+285 VOLTS
2
SRG. SET A
SRG. SET B
12R
13R
14R
15R
TUBE E
TO TUBE H
3
+180 VOLTS
MAXIMUM SPEED ADJUSTMENT
16R
1PT
INC.
10C
8C
7C
9C
GEN. A
GEN. B
TACHOMETER GENERATOR A
20R
17R
TUBE F
IR DROP
+105 VOLTS
4
2PT
MINIMUM SPEED ADJUSTMENT
3PT
SPEED CONTROL (MOTOR DRIVEN)
4PT
VERNIER SPEED ADJUSTMENT
0 VOLTS
5
DC CURRENT BALANCE ADJUSTMENT
IR DROP
(b)

Figure 7. D-c balance control circuit for parallel operation of both converter sets

Figure 8. A-c balance control circuit for parallel operation of both converter sets


VOLTS/CYCLE CONTROL POTENTIOMETER
VOLTS/CYCLE CONTROL CIRCUITS
PHASE-SHIFTING BRIDGES
RECTIFIERS
ALTERNATOR A
ALTERNATOR CONTACTOR
CONVERTER A
ALTERNATOR B
ALTERNATOR CONTACTOR
CONVERTER B
IR DROPS PROPORTIONAL TO FIELD CURRENTS
1-2 VOLTS/CYCLE
1-4 VOLTS/CYCLE
PC
(a)
SET A
SET B
+255 VOLTS
+180 VOLTS
+105 VOLTS
0 VOLTS
2
3
4
5
6RA
24R
TUBE L
TUBE M
TUBE S
TACHOMETER GENERATOR A
FIELD OF ALTERNATOR A
FIELD OF ALTERNATOR B
53R
54R
27R
35R
34R
14C
7C
8C
13C
11C
7PT
BIAS ADJUSTMENT
11PT
RANGE ADJUSTMENT
9RC
8PT
VOLTS PER CYCLE CONTROL
4X REACTOR
IR DROP PROPORTIONAL TO FIELD CURRENT
12PT
A-C CURRENT BALANCE ADJUSTMENT
IR DROP PROPORTIONAL TO FIELD CURRENT
(b)
SECONDARY OF TRANSFORMER WHOSE PRIMARY IS CONNECTED ACROSS ONE LEG OF CONVERTER SECONDARY DOUBLE-DELTA WINDING THUS:

with respect to the field current of alternator *A*.

Stabilizing Circuits

The complete system uses six regulators to control the d-c generator and alternator fields. These are the following:

1. Frequency regulator (tachometer generator signal compared to an adjustable reference voltage).
2. Current limit regulator (a voltage proportional to d-c loop circuit current compared to an adjustable reference voltage).
3. Volts per cycle regulator (a direct voltage proportional to converter voltage compared to tachometer generator voltage).
4. Maximum converter voltage (a direct voltage proportional to converter voltage compared to an adjustable reference voltage).
5. D-c balance-parallel operation (d-c loop circuit current of set *B* compared to d-c loop circuit current of set *A*).
6. A-c balance-parallel operation (field current of alternator *B* compared to field current of alternator *A*).

As in many regulators, these circuits tend to have a regular hunting frequency under normal running conditions. Thus it is necessary to introduce stabilizing circuits to eliminate oscillations under steady state operating conditions. These circuits are chosen to give fast response to transient changes as well as optimum stability under steady state conditions.

The dotted portion of Figure 4b represents the stabilizing elements for the frequency control circuit. A capacitor and resistor connected from plate to grid of tube *F* feeds a signal back to the grid of the tube, which is leading in phase angle with respect to any steady oscillating voltage from plate to cathode of the tube. Such a circuit reduces the gain of the amplifier with respect to the a-c hunting component of grid voltage without sacrificing d-c or steady state sensitivity. The circuit also has the characteristic of attenuating any high frequency ripple voltage on tube *F* grid caused by tachometer generator ripple or stray voltages induced in the leads. In general, the practical effect of the plate to grid stabilizing circuit is to slow down the transient response of the system. To offset this, a resistor capacitor feedback from the positive d-c generator lead to the tube *F* grid is used. Once a suitable combination of resistance and capacitance for plate feedback is found by trial, a generator voltage feedback circuit can be found which will eliminate steady state oscillations and provide the transient response desired.

The current limit circuit is sufficiently stable without the use of stabilizing circuits. The current limit amplifier has a much lower gain than the amplifier of other circuits, which accounts for the inherent stability of this circuit.

In Figure 6b, four stabilizing circuits are shown, two for the volts per cycle regulator and two for the maximum converter voltage regulator. In each case the plate-to-grid circuit is combined with the alternator field voltage feedback circuit to produce maximum speed of response and stability. In this case, since the alternator field is energized from a controlled thyratron rectifier, a resistor capacitor filter is used to produce a smooth d-c voltage for use as a feedback signal.

It is to be noted that the stabilizing circuits have little or no effect on the accuracy of the system. Under normal running conditions, the various voltages in the vacuum tube circuits are d-c voltages; hence the stabilizing capacitors charge up to the voltage impressed on them and remain at that charge until the voltage changes. Thus the gain of each circuit and the established adjustments are unaffected by the stabilizing circuits.

Control Panels and Console

Contactors, relays, electronic regulators, knife switches, and transformers, required to control the power system, are mounted on two enclosed dead front switchboards located in the machinery room with the converter sets (see Figure 9). Switchboard *A* contains the starter for the synchronous motor and all other controls for the constant speed and the converter sets, including the electronic regulators (see Figure 10). Switchboard

B controls the output circuits from the converters and includes all devices used in conjunction with the utilization and metering channels.

Pilot control switches, indicating lights, and instruments, which the operators require to control the system, are arranged on the console which is located in the control room above the machinery room, level with the wind tunnel test chamber (see Figure 11). The console is of semicircular design with an elongated center section. The right half controls the wind tunnel itself, and the left half controls the power model testing equipment (see Figure 12).

The console is so arranged that the operator in charge of power model testing can read all instruments and operate all control switches from one position. Instruments are mounted on a raised inclined section. Control switches and indicating lights are mounted on the horizontal section, which also provides writing space for keeping test records. The back is sufficiently low so that the operator can observe the wind tunnel test chamber through a window.

Figure 9 (left). General view of motor generator set room with switchboard *A* in background and switchboard *B* on left side

Figure 11 (right). Control room with operators seated at console, tunnel test chamber visible through window in background

Operation

Before a power model test is started, the operator closes the knife switches to the channels to which power model motors are connected; he closes the converter range switches in the position for proper volts per cycle. Then he starts the constant speed motor generator set from the machinery room. When the "ready" light on the console glows, indicating that the thyratron tube filaments are at operating temperature, full voltage is applied to the d-c motor fields, the alternator and d-c generator field circuits to the electronic regulators are closed. The operator then takes over from the console and goes through the following steps:

1. With the motor operated speed rheostat in the position for minimum converter speed, and the volts per cycle rheostat in the minimum voltage position, the operator closes the bus energizing switches, thereby closing the a-c line contactors and the d-c

Figure 10 (left). Section of switchboard *A* containing electronic control equipment

Figure 12 (below). Left section of control console containing instruments and control switches for power model testing equipment

loop circuit contactors, and thus starting the converter sets. The converter sets turn at minimum speed.

2. Volts per cycle is adjusted by turning the rheostat up, until volts per cycle corresponds to the rating of the model under test.

3. When model motors are tested from standstill, the operator now closes control switches, which in turn control contactors (1*A* to 6*A*) connecting the selected channels to their bus. When model motors are tested in the tunnel and their propellers are windmilling, the operator first raises the frequency to a value corresponding to the windmilling speed, so that power is applied to the motors at a low slip.

4. Model motor speed then is adjusted to the desired value by adjusting the system frequency. A control switch permits raising and lowering the motor operated rheostat, and a manually operated rheostat serves as a vernier adjustment.

5. In order to meter any channel, the operator closes the metering switch of that channel. This causes the *B* and *C* contactors of that channel to close and the *A* contactor to open. Next, the operator turns the transformer ratio selector switches until the pilot voltmeter and ammeter are reading up scale. Closing a test switch connects the precision wattmeter in the metering circuit.

6. Any or all channels are de-energized by tripping out the busses, since the channel contactors cannot interrupt power. If an individual channel is to be dropped during a test, a push button may be pressed which reduces volts per cycle to zero. The channel contactor then can be tripped, and, when the push button is released, voltage returns to the previously set value.

When the two converter sets are to be operated in parallel, the paralleling contactor is closed from the console with the converter sets at standstill. The secondaries of the converter sets are connected in parallel. When starting the converter sets, the operator momentarily raises volts per cycle, and the two converter sets lock in step. Frequency and volts per cycle controls for set *A* control both sets during parallel operation.

Conclusions

The control system described in this paper realizes the following operating advantages:

1. Accurate control of system frequency permits accurate adjustment of model motor speed.

2. Accurate volts per cycle control over the frequency range prevents damage to model motors on account of voltage drift during a test.

3. Current limit and maximum voltage controls prevent damage to the rotating machines.

4. Splitting the system into two sections provides flexibility, as either two tests with half system power or one test with full system power may be scheduled. Also two independently adjustable power supplies are available for studies on counter rotating propellers.

5. During parallel operation, the electronic control balances the two converters with respect to load and power factor, thereby freeing the operator from continually readjusting the two converter sets by hand.

6. Providing several channels on one power bus and one metering channel per bus permits precision metering of several model motors with the investment of only one set of precision metering equipment.

Reference

1. Precision Speed Measurements for Wind Tunnels, R. K. Fairley, H. L. Clark. *Transactions*, American Society of Mechanical Engineers, volume 68, October 1946, pages 719–22.

Modern Excitation Systems for Large Synchronous Machines

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AN EXCITATION SYSTEM consists essentially of an exciter and means for its regulation. Originally, manual control of voltage was used widely. With the development of power systems using modern apparatus, automatic control, in general, has come to play an important part in the progress of the electric power industry. Therefore, it is in keeping with this progress that voltage regulators have been used increasingly and have become an important factor in making possible further economies and improvements in system reliability. The increased use and acceptance of voltage regulators has been a great stimulus to the advancement of the art with the result that with the past few years several new types have been developed, placed in regular service, and are performing successfully.

The need for automatic regulation of voltage for water wheel generators and synchronous condensers has long been recognized. One of the important factors which has directed attention recently to the problems of automatic regulation of excitation has been the desirability of obtaining increased capacity in large 3,600-rpm turbine generator sets. The physical size of the generating element for a given capacity is determined largely by its short-circuit ratio. Reduction in short-circuit ratio makes more desirable the use of automatic regulation, and at the same time increases the importance of the excitation system reliability. With this a desire also has come to provide sources of excitation other than the direct-driven exciter for the higher speed units.

This paper reviews the principles and characteristics of existing widely used excitation systems and discusses some of the more recently developed excitation systems. These systems and the solutions of the general excitation problem discussed in this paper are identified primarily with the company with which the authors are associated. Any terms used are based on either American Standards Association definitions or on generally accepted industry terminology.

Types of Excitation Systems

The present day practice in excitation systems is to use the unit system which consists of an individual exciter and regulator for each a-c machine. In this general type of system, the voltage of each a-c machine is automatically and independently controlled by its own regulator, and the division of reactive current among machines is automatically controlled by alternating current compensation of the individual regulators. There is no problem of load division among exciters, as each exciter is operated and controlled independently. Each of the excitation

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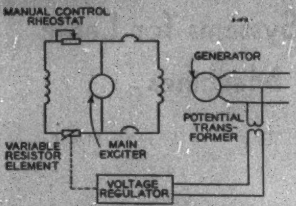


Figure 1. Elementary diagram of an excitation system consisting of a main exciter and a direct-acting rheostatic-type voltage regulator

systems discussed is designed to operate as a unit system.

Self-Excited Main Exciter and Direct-Acting Rheostatic-Type Regulator

One of the simplest types of excitation systems consists of a self-excited main exciter and a direct-acting rheostatic-type regulator.[3] The essential elements of such a system are shown schematically in Figure 1. The exciter is a shunt wound d-c machine and supplies excitation to the field of the main synchronous machine through a field breaker. In normal operation the voltage regulator detects changes in the controlled a-c voltage and mechanically changes the resistance in the exciter field circuits to produce the required exciter voltage. This regulator can be taken out of service and the exciter placed on manual control of voltage by using the rheostat provided for this purpose.

Main and Pilot Exciters and Indirect-Acting Rheostatic-Type Voltage Regulator

An excitation system which has been used with the larger machines for the past few years consists of a main and pilot exciter and an indirect-acting rheostatic-type regulator. The essential elements of such a system are shown schematically in Figure 2. The main exciter is a separately excited d-c machine and supplies excitation to the field of the main synchronous machine through a field breaker. The pilot exciter is a compound-wound self-excited d-c machine designed to hold essentially constant voltage. The field of the main exciter is energized by the pilot exciter through a motor-operated Wheatstone-bridge rheostat. In normal operation the voltage regulator detects changes in the controlled a-c voltage and adjusts the position of the rheostat to vary the output voltage of the main exciter. This regulator can be taken out of service and the rheostat placed under manual control.

Main Exciter and Static Voltage Regulator With Rotating Amplifier

One of the more recently developed excitation systems consists of either a separately- or a stabilized self-excited main exciter and a static voltage regulator with a rotating amplifier. The essential elements of this system, when using a self-excited main exciter, are shown schematically in Figure 3. The main exciter is a self-excited d-c machine and supplies excitation to the field of the main synchronous machine through a field breaker. In normal operation the voltage regulator detects changes in the controlled a-c voltage and adjusts the output of the rotating amplifier to either decrease or increase the exciter field voltage, which in turn controls the output voltage of the main exciter. The static voltage sensitive element can be of either the impedance or the electronic type. The rotating amplifier can be taken out of service and the exciter placed on manual control by using the rheostat provided for this purpose.

This system was developed after several installations of electronic regulators and electronic exciters on hydrogen-cooled synchronous condensers indicated that the cost of the electronic equipment would be considerably more than that of a standard excitation system with rotating exciter. Studies indicated that the performance of the electronic system could be duplicated by the static regulator and rotating exciter, which has a cost comparable to the standard excitation system. Experience over a period of five years has amply confirmed the expected performance.

Electronic Exciter and Electronic Voltage Regulator

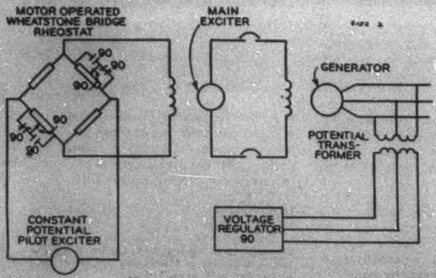
Another type of excitation system consists of an electronic exciter and an electronic voltage regulator. The essential elements of such a system are shown schematically in Figure 4. The electronic exciter is a power tube rectifier, complete with its required control equipment and supply transformer, and supplies excitation to the field of the main synchronous machine through anode breakers which act as a field breaker. In normal operation, the voltage regulator detects changes in the controlled alternating voltage and adjusts the firing time of the tubes to give the required exciter output voltage. Provision also is made for operating the electronic exciter on hand control.

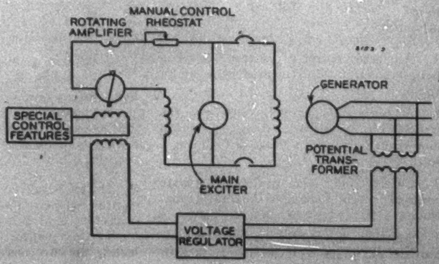
Exciters

The exciter for any large synchronous machine must meet certain minimum performance requirements. These requirements have not been formulated specifically on an industry wide basis, but practice and experience have demonstrated

Figure 2 (below left). Elementary diagram of an excitation system consisting of a main and pilot exciter and an indirect-acting rheostatic-type voltage regulator

Figure 3 (below right). Elementary diagram of an excitation system consisting of a main exciter and a static voltage regulator with a rotating amplifier


MOTOR OPERATED WHEATSTONE BRIDGE RHEOSTAT
90
MAIN EXCITER
GENERATOR
POTENTIAL TRANS-FORMER
CONSTANT POTENTIAL PILOT EXCITER
VOLTAGE REGULATOR 90



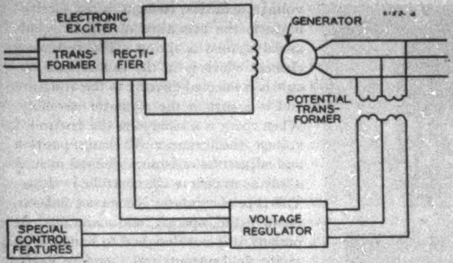


Figure 4. Elementary diagram of an excitation system consisting of an electronic exciter and an electronic voltage regulator

what they should be. As applied, present day exciters have conservative current and voltage ratings and a reasonable ceiling voltage. The over-all excitation system is designed to have a moderately quick response. The foregoing requirements have been arrived at largely through experience with shaft-driven exciters, and, in cases where other than shaft-driven exciters are used, the excitation system should be designed to give the same actual performance under certain assumed adverse system operating conditions.

SHAFT-DRIVEN EXCITERS

Shaft-driven exciters have been generally used for many years with turbine generators, water wheel generators, synchronous condensers, and frequency changers. The outstanding advantage of a shaft-driven exciter is its ability to ride through severe system disturbances while continuing to maintain the excitation. The main disadvantage of a shaft-driven exciter is that maintenance must be done either by shutting down the main machine or by incurring the hazard of maintenance while the machine is operating.

HYDROGEN-COOLED EXCITERS

In the past motor-driven exciters were generally accepted for use with hydrogen-cooled synchronous condensers and frequency changers. This practice allowed the main machines to be built with a simple hydrogen-tight shell with no running-shaft hydrogen seals.

During the last five years, successful experience has been obtained with direct-connected main exciters on hydrogen-cooled synchronous condensers operating up to speeds of 900 rpm. In each of these installations, the exciter operates in a hydrogen-filled compartment which is a part of the main shell. When the machine is at rest, the exciter compartment can be isolated from the main shell and opened for inspection and maintenance. The use of a direct-connected pilot exciter in these applications would have required major structural changes and a longer shaft extension, which would have materially increased the cost.

SPECIAL MOTOR-DRIVEN EXCITERS

The increased use of high-speed turbine generators of larger capacity has accentuated some of the problems of the shaft-driven exciter. These problems have been largely met by both improvements in design of high-speed exciters and also by the use of geared exciters for the larger sizes. However, with the trend during the past few years toward still larger capacity high-speed turbine generators, there has been an increasing inclination to provide excitation from sources other than shaft-driven exciters. A special motor-driven exciter has been developed primarily for use with large high-speed turbine generators, and also may find more general application for other types of synchronous machines. There are some obvious advantages for such a motor-driven exciter set.

1. The foundation of the main turbine generator set is simplified.

2. A low speed commutator is obtained without the use of gears.

3. Maintenance can be readily accomplished without shutting down the main unit when a spare exciter is provided.

Such a motor-driven exciter can obtain its power from the station auxiliary bus and be considered as one of the essential station auxiliaries. When the station auxiliary power is provided by either shaft generators for separate house generators, the exciter set does not need to be particularly special in design. However, most steam electric stations obtain auxiliary power either from the main generator terminals or from the main station bus, and the auxiliary power system is subjected to reduction in voltage for all main system disturbances. It is essential, therefore, to design the motor-driven exciter set to carry on through such system disturbances. It has been shown[5] that it is possible to design a motor-driven exciter set which gives the same main generator and system performance as is obtainable with a direct-connected exciter for sustained system voltage as low as 70 per cent of normal and severe faults of 12 to 15 cycles duration.

The essential features of such a motor-driven exciter set are as follows:

1. The inertia of the set is increased by using a flywheel located between the motor and generator. An inertia constant of five based on the exciter nameplate kw rating is conservative.

2. The induction motor should have a high maximum torque to enable it to carry the output of the exciter at ceiling voltage. This will be accomplished by using a motor with a maximum torque approximately equal to five.

3. The exciter should have a normal ceiling sufficient to realize a ceiling of not less than 120 per cent when the driving motor is operating, with approximately 70 per cent voltage applied to its terminals.

EXCITER STABILITY

The ability of the exciter to maintain its voltage without appreciable drift with time has been actively discussed[14] as a requirement for the manual control of rotating self-excited exciters. When automatic voltage regulators are in use, such a requirement is not particularly pertinent. Even when under manual control, such a requirement is not necessary for short periods of operation such as transfer from one source to another. It may be important, however, when the a-c generator is left unsupervised under manual

Figure 5. Electronic exciter for a synchronous condenser

control of its excitation for prolonged periods, particularly at light loads.

Electronic Exciters

The electronic exciter offers another solution for the elimination of the direct-connected exciters. Electronic exciters offer similar advantages to those already outlined for motor-driven exciters.

Electronic exciters have been supplied for differing applications. Three large synchronous condensers using electronic exciters and supplied by one manufacturer have been in operation each for 11, 9, and 7 years, respectively. One of these electronic exciters is shown in Figure 5. Functionally, these electronic exciters with their electronic regulators have given an outstanding performance. They are equipped with water-cooled ignitron power tubes having ample capacity and arranged for changing tubes while the set is in operation. Although initially the tube life did not measure up to expectations, progress in the art has resulted in reasonable tube life. Based on present experience, such an electronic exciter obtaining power through a suitable transformer from the auxiliary source in the station will have a considerably higher initial cost and higher maintenance than either the shaft- or motor-driven exciters.

Some users have insisted that the supply for an electronic excitation source should be a shaft-driven a-c generator on the main set. This brings back most of the structural disadvantages of a direct-driven excitation source and makes such a solution even more expensive.

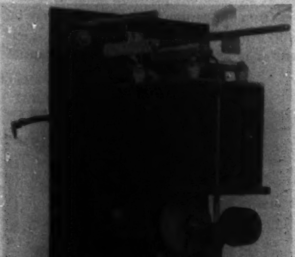


Figure 6. Direct-acting rheostatic-type voltage regulator

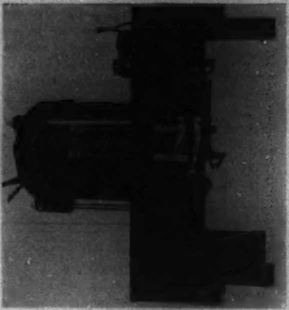


Figure 7. Indirect-acting rheostatic-type voltage regulator

When the power for an electronic exciter is obtained from the main alternator terminals or the system, it is important to design the transformer supplying the electronic exciter with a turn ratio such that high ceiling voltages are available. This is necessary for two reasons.

1. To realize at least 120 per cent ceiling with 70 per cent alternating voltage available.
2. To realize a reasonable response during a system disturbance.

The experience with electronic exciters on synchronous condensers has demonstrated that they can be applied successfully with other types of synchronous machines.

Voltage Regulators

In recent years, the use of voltage regulators has increased considerably. The main function of these regulators is to respond to any changes in the controlled voltage and to instigate the necessary changes in the excitation system to restore the controlled voltage to its original value. The required performance of the voltage regulator can be stated quite simply in terms of its band of voltage regulation which, broadly speaking, includes dead band, inherent regulation, accuracy, and stability. The various types of regulators which have been developed employ different basic principles, as will be described.

Direct-Acting Regulators

One of the simplest voltage regulators is the direct-acting rheostatic type,[2] as shown in Figure 6. In this regulator the voltage sensitive element is an electromagnet, the armature of which is balanced against a spring. The resistance element which is in the exciter field circuit is connected directly to the armature and is a part of the regulator assembly. When there is a change in the controlled voltage, the armature will change position and adjust the resistance element in such a way as to restore the controlled voltage. This type of regulator is compact and simple to install, operate, and maintain. At present, it is not designed to handle high exciter field currents and, therefore, its use is limited to small and medium size machines. This regulator is of the continuously-acting type and has a narrow band of voltage regulation.

Indirect-Acting Regulators

The indirect-acting rheostatic-type voltage regulator is well suited for use in excitation systems having both a pilot and a main exciter and is shown in Figure 7. The voltage sensitive element of this regulator may be a polyphase torque motor, the torque of which is proportional to the approximate average 3-phase voltage. The torque of the motor is balanced against the pull of a spring so that for each value of voltage there is a definite position of the motor rotor to which is attached a contact assembly. The contact assembly has two sets of contacts, the first set operates intermittently for small changes in voltage to control the pilot motor of the motor-operated Wheatstone-bridge rheostat (Figure 8), which in turn adjusts the resistance of the rheostat. Both sets of contacts operate for large changes in voltage, the second set controlling the temporary switching of all the regulating resistance in the rheostat, while the first set of contacts controls the final position of the rheostat. Thus, for large changes in voltage, there is an extremely rapid change in the excitation field resistance which increases the response of the entire excitation system.

This type of regulator is rugged and reliable. When used with the Wheatstone-bridge-type rheostat, rapid response is obtained, and practically no limit need be placed on the field currents it will successfully control. For some applications where severe fluctuating loads are encountered, the dead-band principle in conjunction with the intermittent control feature, is not the ideal type of regulator.

Impedance Regulator

The impedance regulator is one of the more recent developments in the voltage regulator field and is shown in Figure 9. It is a static-type regulator and is normally used with a rotating amplifier. The function of the rotating amplifier is to respond to changes in the controlled voltage as detected by the regulator and modify the voltage in the field circuit of the main exciter in such a manner as to restore the controlled voltage.

The impedance voltage regulator consists essentially of a linear and nonlinear impedance. The 3-phase voltage to be controlled is passed through a positive-sequence network and the resulting single-phase voltage, which is the approximate average of the 3-phase voltage, is impressed on both the linear and nonlinear impedances. The linear impedance supplies rectified current to the control field of the rotating amplifier in such a direction as to raise the output voltage. The nonlinear circuit also supplies rectified current to the control field of the rotating amplifier, but in such a direction as to lower the output voltage. The volt-ampere characteristics of the two impedances are such that they intersect at some value of control voltage at which point there is no net current in the control field of the rotating amplifier. For a change in the controlled voltage, there will be a net current in the control field which is in the direction to restore the controlled voltage to its previous value.

The impedance voltage regulator is simple in operation, has no moving parts, and no inherent dead band, and is of the continuously acting type. Special control features can be added readily to this type of regulator by the use of additional control fields in the rotating amplifier. The use of this type of regulator results in fast smooth response of the excitation system with a narrow band of voltage regulation.

Electronic Regulator

Another static type of voltage regulator is the electronic regulator which is shown in Figure 10. It can be used either with a rotating amplifier in an excitation system with rotating exciters or directly with an electronic exciter to control the grids of the firing tubes, which in turn control the firing point of the main power tubes.

The electronic regulator consists essentially of a constant voltage glow tube and the necessary rectifier and control tubes. The 3-phase voltage is rectified and compared with the voltage across the glow tube. Any difference in these voltages is amplified and produces a change in the grid of the tubes supplying power to the control fields of the rotating amplifier. The tubes used are radio type, and, to give reliable operation, two tubes in parallel are used so that tubes can be checked or changed without shutting down the regulator.

The electronic regulator has the same features as the impedance regulator. In this type of regulator, special control features can be added directly to the regulator. Its use also will result in fast smooth response[1] of the excitation system with a narrow band of voltage regulation.

The outstanding performance of several installations with electronic exciters and electronic regulators, in maintaining smooth voltage control at locations where severe fluctuating loads exist, was caused primarily by the characteristics of the regulator and not by the electronic exciter. It is well recognized that a voltage regulator of the continuously acting type andwithout dead band gives the ideal type of control. These regulating characteristics are now available for using with rotating exciters.[1]

General

Rheostats

Pilot exciter field rheostats are usually of the manually-operated dial type. They are used initially to obtain the proper pilot exciter voltage and once that is done need not be changed.

Figure 8. Motor-operated Wheatstone-bridge rheostat

Main exciter field rheostats should be chosen to suit the application. In general, they will be motor-operated so that they may be controlled from a central location. When the main exciter is operated self-excited with a voltage regulator, the rheostat will be used for short intervals only while maintenance work is being done. In this case, a relatively simple and inexpensive dial type rheostat may be used.

When the main exciter is excited separately and the motor-operated Wheatstone-bridge-type rheostat is used in conjunction with a rheostatic-type voltage regulator, such a rheostat is of the face-plate type as shown in Figure 8. This rheostat is simple and rugged and able to successfully withstand the frequent operations required by the regulator.

Several years ago, it was considered general practice to use main generator field rheostats. There were two reasons for requiring these rheostats.

1. More than one synchronous machine often was excited from a common d-c source and the rheostat equalized any difference in the excitation requirements.

2. The excitation systems were not designed for operation at the low voltages required for line charging and for synchronous generators with low short-circuit ratios.

In modern practice, the unit system of excitation is used almost universally and the entire excitation system is capable of operating down to low voltages. Thus the need for main generator field rheostats has practically disappeared.

Dynamic Stability

The possibility of operating above the ordinary steady state stability limits by using a special excitation system has been recognized for some time, although it has not been ordinarily necessary to require operation in this region. This requirement together with equipment which is capable of so operating may be desirable during emergency conditions in order to allow for safe operation with smaller stability margins and, therefore, with higher transmission system loadings. An interesting application of this principle has been made to electric ship propulsion.[12]

Reference 12 shows, in general, the factors which need to be taken into account in order to provide for stable operation. A desirable characteristic is small voltage dead band with small excitation system time lags.

Lower Excitation Limit

If the unit is under automatic control of voltage, it may be desirable to use a fixed or variable lower limit. A variable limit for generators preferably should be made a function of prime mover torque, which can be obtained conveniently from the position of the mechanism which controls prime mover input. Such a variable limit allows for a greater range of regulator operation than if a fixed lower limit is used and insures against the possibility of reducing the excitation below the point where instability may result.

The critical value of excitation, below which instability may result, does not depend greatly upon the external system. It is, therefore, possible by means of a variable lower limit to insure an adequate amount of excitation for a wide range of system conditions. This lower limit is insurance against an error in setting the voltage adjusting rheostat or other errors which incorrectly may reduce the excitation.

Salient-pole synchronous condensers may be provided with a fixed lower limit of negative polarity and possibly biased by terminal voltage in order to take advantage of the saliency torque, and thereby increase the lagging capacity.[13] Frequency changer sets, however, may require an adjustable lower limit which can be adjusted as determined by operating conditions.

A fixed lower limit of generator excitation is often used. With the use of a low short-circuit ratio generator, this lower limit necessarily must be set at a value higher than that required for synchronizing. Therefore, during synchronizing the limit must be removed or synchronizing obtained under manual control of the excitation.

UPPER EXCITATION LIMIT

An upper excitation limit has been used for some applications based on limiting either the excitation voltage, alternator field current, or armature current. A common method in supervised stations is to ring an alarm when either the excitation voltage or current exceeds a given amount for a predetermined time. Also, for synchronous condenser installations, a limit on the excitation has been based on the average field temperatures (from resistance measurement) exceeding that obtained from rated or normal field excitation. It should be noted here that such a limit should not be based on allowable temperature for insulation used, but rather should be based on the total temperature obtained with nominal or rated excitation and the existing ambient. Although many variations of the use of an upper limit are possible, it probably is necessary only to sound an alarm in supervised stations when the excitation exceeds for a given time delay a limiting amount. In unsupervised stations it may be desirable to limit the excitation to the rated value and allow for a corresponding drop in voltage at the regulated bus. This need be the only requirement necessary for 0.8- to 0.9-power-factor generators since their rated field excitation will be normally sufficient to carry the maximum kw output of the prime mover.

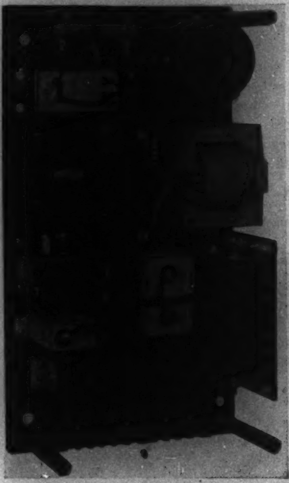


Figure 9. Impedance-type voltage regulator

Application Guides

The following discussion summarizes what may be considered modern practices for successful excitation systems of the larger synchronous machines. In present day practice, rotating exciters are almost exclusively used, and therefore, the preferred field of application of electronic exciters will be left until such time as they can be justified on an economical basis and until they are more generally accepted. The unit system of excitation, that is an independent exciter and regulator for each synchronous machine, is assumed throughout.

TURBINE GENERATORS

For several years it has been considered standard practice to control and supply the excitation of the larger turbine generators with a main exciter, pilot exciter, Wheatstone-bridge rheostat, and indirect-acting-type voltage regulator. This system has proven to be a reliable one, but experience has indicated that there are some fundamental disadvantages. The inherent dead band requires greater droop to obtain accurate division of reactive current between machines, which in turn required more frequent adjustment of voltage levels to maintain voltage at some desired location on the system. Special control features cannot be added readily to this type of regulator.

One other solution for the small and medium sized generators has been to apply the direct-acting-type regulators. In this way the pilot-exciter and Wheatstone-bridge-type rheostat are eliminated. The continuous-type regulator gives good division of reactive current and smooth response. The regulator has the disadvantage that at present it cannot handle high exciter field currents, and, therefore, its use is limited to generators not in excess of 25,000 kva. It is also difficult to add special control features to this type of regulator.

Experience indicates that a universally applicable regulating system for the larger turbine generators is a static-type regulator with a rotating amplifier. Smooth fast response is obtained; the continuous-type regulator with no dead band gives accurate division of reactive current, and low energy control circuits make it possible to add automatic limits. In addition, this type of regulator is readily applied to existing main exciters and in this way any excitation system can be modernized conveniently.

For those who still prefer shaft-driven exciters for their high speed units, present day conservative practice would suggest gear-driven exciters for the larger units and direct-driven for the smaller units up to 25,000 kva. However, with the improvements in system relaying and auxiliary power system arrangements, the possible simplifications, flexibility, and performance resulting from the use of a special motor-driven exciter make its use particularly attractive for large turbine generators. It is expected that the acceptance of motor-driven exciters in these applications will increase rapidly.

WATER WHEEL GENERATORS

It is almost universal practice on all of the larger water wheel generators to control and supply their excitation with a main exciter, pilot exciter, Wheatstone-bridge rheostat, and indirect-acting-type voltage regulator. Hydro stations usually feed radially into a system so that any disadvantage of the inherent dead band in the regulator is not very important. The high speed feature of the bridge rheostat makes it ideal for controlling voltage during loss of load, and the ability to operate at very low excitations makes it ideal for line charging requirements.

Because of the additional advantages of the static-type regulators and a rotating amplifier, it can be expected that this type of regulating system eventually will be applied to water wheel generators.

Figure 10. Electronic-type voltage regulator

Direct-connected exciters are used almost universally in hydro installations, and maintenance has not proved excessive in these slow speed exciters.

SYNCHRONOUS CONDENSERS

Several years of experience with static-type regulators and rotating amplifiers on synchronous condensers, indicates that this is practically the ideal regulating system for such applications. The continuous-acting-type regulator with no dead band makes it particularly suited to the type of voltage control expected of synchronous condensers.

During this same time, direct-connected hydrogen-cooled exciters have been operating successfully with several synchronous condensers. As a result of this experience, all moderate and low speed synchronous condensers, both air-cooled and hydrogen-cooled, may now be equipped with direct-connected exciters.

FREQUENCY CHANGERS

The requirements of an excitation system for a frequency changer set will vary between that of a synchronous condenser and water wheel generator, depending upon the location of the set on the system. Most of the existing frequency changers are equipped with main and pilot exciter Wheatstone-bridge rheostats and indirect-acting voltage regulators. This type of excitation system has given successful performance. Where system conditions require close control of voltage, a static-type regulator and rotating amplifier can be used to advantage.

As a result of the successful experience with hydrogen-cooled exciters on synchronous condensers, all frequency changers, both air-cooled and hydrogen-cooled, now may be equipped with direct-connected exciters.

Conclusions

1. The general use of automatically controlled excitation systems makes it possible to operate existing equipment at higher loadings and to design more economical system extensions.

2. Improved division of reactive currents and improved system performance are made possible by applying voltage regulators, which have a narrow band of voltage regulation now available in modern regulators of the continuously acting type.

3. The advantages of providing limits with voltage regulators are recognized, and these limits may be readily applied when static-type voltage regulators are used.

4. Excitation systems which do not include pilot exciters may be readily modernized by applying a rotating amplifier and a static-type voltage regulator.

5. Special motor-driven exciters may be successfully used in place of large high speed shaft-driven exciters.

6. Shaft-driven hydrogen-cooled main exciters have been successfully used for moderate speed synchronous machines.

7. Co-ordinated excitation systems are now available for all types of application requirements.

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High Vacuum Leak Testing With the Mass Spectrometer

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THE DETECTION of small leaks in large all-metal vacuum systems has long been a troublesome manufacturing problem for which there was no satisfactory solution. All of the generally used techniques of leak detection are cumbersome or have serious limitations such as lack of sensitivity, lack of selectivity, poor response, dependence upon vacuum conditions, uncertainty of interpretation of response, or usefulness being restricted to certain materials. Experience with many methods of leak detection on all sizes of apparatus has shown that an ideal leak detector must have the following characteristics:

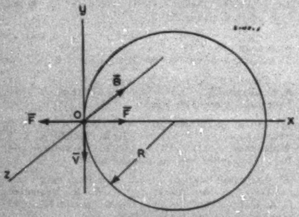

Figure 1. Forces acting on a charged particle in a magnetic field

(*a*). *High sensitivity.* The instrument must be able to detect small leaks.

(*b*). *Accuracy.* To facilitate repairs to the system the leak must be accurately located.

(*c*). *Response.* The instrument must respond quickly when a leak is located and must "clean up" quickly when the leak has been passed. Too much lag interferes with the accurate location of leaks.

(*d*). *Selectivity.* The leak detector must be capable of locating large or small leaks in the presence of many other leaks.

(*e*). *Certainty of response.* The instrument must always respond to a leak and must not be subject to spurious effects or false signals.

(*f*). *Versatility.* The method should be applicable to large or small systems constructed of metal, glass, or other materials. Usefulness of the method should not depend too much upon vacuum conditions, since it is usually not convenient to obtain a high vacuum in the test sample.

(*g*). *Simplicity.* Operation of the leak detector must be sufficiently simple that it can be performed by unskilled personnel.

(*h*). *Portability.* Since many vacuum systems are permanent installations, the leak detector must be portable.

It is the object of this paper to describe a special mass spectrometer, designed for use as a leak detector, which fulfills these requirements to a higher degree than any previously known form of leak detector.

This leak detector design is based on the mass spectrometer experience of Doctor A. O. C. Nier of the University of Minnesota,[1] who also shared in a major way in the problems of design and manufacture of the leak detector.

The Mass Spectrometer

The mass spectrometer is a device which utilizes electric and magnetic fields to sort out ionized gas molecules according to their molecular weights (or to be exact, according to mass-to-charge ratio, or *specific mass*). The instrument is therefore capable of measuring the partial pressure of each gas in a mixture of gases in a vacuum system. The device is thus inherently selective, which immediately suggests its use as a detector of some gas which can be used as a tracer in a mixture of gases. In the present application, the tracer gas would be allowed to flow through a leak, and its presence in the vacuum system would be detected by the mass spectrometer, irrespective of the presence of other gases. It remains to determine whether the other requirements of an "ideal leak detector" can be met.

Theory

In order to understand the principle of operation of the mass spectrometer, let us consider the forces acting on a charged particle moving at right angles to a magnetic field as shown in Figure 1. In vector notation, the electromagnetic force acting on the particle will be

$$\mathbf{F} = q\mathbf{v} \times \mathbf{B} \qquad (1)$$

where $\mathbf{F}$, $\mathbf{v}$, and $\mathbf{B}$ are vectors representing force, velocity, and magnetic induction respectively, and q is the charge on the particle.

Since the force is always perpendicular to the velocity, there will be no linear acceleration; but there will be constant angular acceleration, so that the path of the particle will describe a circle in the x-y plane with the center at $(R, 0, 0)$ as shown, where R is the radius of curvature.

From elementary mechanics, a particle of mass m moving with a velocity v along a curved path of radius R will experience a centrifugal force

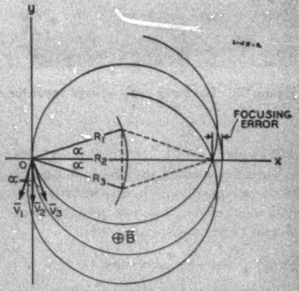

Figure 2. Focusing property of magnetic field

$$F' = \frac{mv^2}{R} \qquad (2)$$

Since F and F' are equal in magnitude but opposite in direction, we can write

$$qvB = \frac{mv^2}{R}$$

or

$$qB = \frac{mv}{R} \qquad (3)$$

Let us assume that the charged particle acquired its velocity by falling through an electrostatic potential V, in

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The authors wish to extend credit to Professor A. O. C. Nier, of the University of Minnesota, whose many contributions to the field of mass spectrometry have been of outstanding importance. They also wish to acknowledge the contributions of Doctor R. B. Jacobs in the application of the mass spectrometer to the problems of leak detection. The mass spectrometer type leak detector and methods of application of the instrument were developed by Professor Nier and Doctor Jacobs who also co-operated with the General Electric Company in solving the problems of design and manufacture of the commercial model.

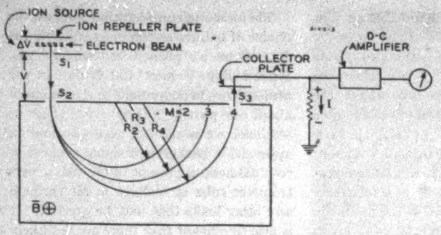

Figure 3 (above left). Mass spectrometer

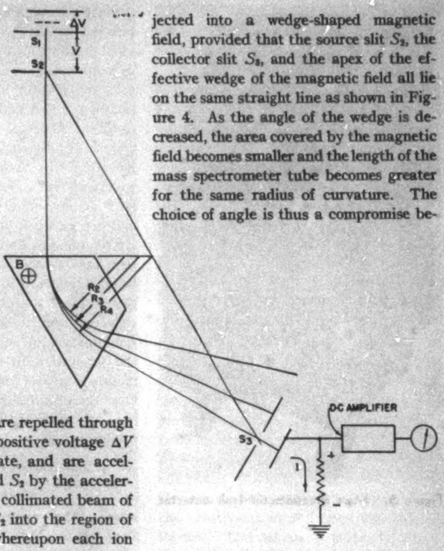

Figure 4 (right). Sixty degree mass spectrometer

which case the potential energy would be

$$PE = qV \tag{4}$$

which must be the same as the kinetic energy after acceleration

$$KE = \frac{mv^2}{2} \tag{5}$$

Setting equation 4 equal to equation 5 and solving for v, we get

$$v = (2qV/m)^{1/2} \tag{6}$$

Substituting this value for v in equation 3, and solving for the mass-to-charge ratio, we obtain

$$m/q = B^2R^2/2V \tag{7}$$

From this we can see that charged particles having different mass-to-charge ratios will follow paths of different radius, and that the radius for any particular particle can be controlled by choosing the proper values of magnetic field and accelerating voltage.

Let us now consider the effect of the magnetic field on charged particles whose velocities at the origin differ in direction by the small angle α, but which are identical in all other respects. A simple geometrical construction as shown in Figure 2 demonstrates that there will be a point, lying on the x axis, at which the three paths are closest together, or at which the originally divergent paths are refocused. The "focusing error" is seen to be

$$\text{Focusing error} = 2R(1-\cos\alpha) \tag{8}$$

where

$$R = R_1 = R_2 = R_3$$

Obviously the focusing error can be made as small as desired by making the angle α small.

The mass spectrometer utilizes the principles that we have discussed in a manner illustrated diagrammatically in Figure 3. Gas molecules are introduced at low pressure to the ion source, where part of the gas is ionized by bombardment with electrons which have been accelerated through a potential of the order of 150 volts. The ions are repelled through the slit S_1 by a small positive voltage ΔV on the ion repeller plate, and are accelerated between S_1 and S_2 by the accelerating potential V. A collimated beam of ions passes through S_2 into the region of the magnetic field, whereupon each ion will follow a circular path according to equation 7. By proper adjustment of the magnetic field and the accelerating voltage, ions of any specific mass

$$M = m/q = B^2R^2/2V \tag{9}$$

can be made to pass through a third slit S_3 and will discharge on the collector plate, causing a positive current to flow through the high resistance to ground. The voltage drop across the resistor is amplified and measured.

If it is assumed that the electric and magnetic fields are adjusted to bring ions of specific mass $M=4$ into focus on the collector plate, ions of specific mass 3 and 2 will follow paths of smaller radius as shown in Figure 3.

In an actual mass spectrometer, since the slits must have a finite width, the beam of ions entering the magnetic field will have a width determined by S_1, and divergence depending upon the geometry of the ion source and slit system. Therefore the collector slit S_3 must be made somewhat wider than the source slit in order to collect all ions of any specific mass. In addition, it is impossible to accelerate all ions through exactly the same electric potential, and the collector slit width must be increased somewhat to collect ions of slightly differing energies.

Up to this point we have discussed only a mass spectrometer in which the path of the ion is bent through a total angle of 180 degrees. It can be shown that the same principles apply when the ions are projected into a wedge-shaped magnetic field, provided that the source slit S_2, the collector slit S_3, and the apex of the effective wedge of the magnetic field all lie on the same straight line as shown in Figure 4. As the angle of the wedge is decreased, the area covered by the magnetic field becomes smaller and the length of the mass spectrometer tube becomes greater for the same radius of curvature. The choice of angle is thus a compromise between minimum tube size and minimum magnet size, and an arbitrary value convenient for manufacturing is generally used. The angle of deflection for the leak detector described in this paper is 60 degrees.

DEFINITIONS AND UNITS

For the purpose of discussing the theory of the mass spectrometer, we have dealt somewhat loosely with such units as mass and charge. In order to facilitate discussion of the special instrument which will hereafter be referred to as the leak detector, it will be useful to redefine the units as follows:

q is always a small integral multiple of the charge of the electron. For the remainder of this paper we will assume that q is one electronic charge
m is the mass of the ion in atomic weight units
$M = m/q$ = specific mass in atomic weight units per electronic charge
B is the magnetic field strength in gauss
V is the accelerating voltage in volts
R is the radius of curvature in centimeters

We now can rewrite equation 7 in terms of these practical units, putting in the constant of proportionality

$$M = 4.8\times10^{-5}B^2R^2/V \tag{10}$$

The Leak Detector

The leak detector is a small portable mass spectrometer of all-metal construc-

Figure 5. Mass spectrometer leak detector

tion designed to detect minute quantities of helium. The instrument continuously draws a sample of gas from the system being leak tested and indicates the partial pressure of helium in the sample. As portions of the test system are exposed to a helium jet, any helium entering the system through a leak causes an increase in the output current of the amplifier, thus indicating the presence of the leak. To insure that it will find very small leaks, the instrument has been designed for the highest sensitivity compatible with ruggedness and simplicity. An over-all view of the leak detector is shown in Figure 5.

The value of the method in actual use will depend upon the ability of the mass spectrometer to detect minute quantities of the tracer gas in a mixture of air and other gases and to detect small changes in the amount of tracer gas.

As an illustration, suppose that we have a vessel with a large number n leaks of equal size. We wish to locate each leak accurately by the use of our leak detector. If we completely cover one leak with the probe gas and leave the remainder exposed to air, the ratio of the flow of probe gas into the vessel to the flow of air is $1/(n-1)$ (assuming equal rates of flow for the different gases). In order to detect the leak, we must admit a detectable quantity of helium to the leak detector, and at the same time we must be able to admit $(n-1)$ times as much air without causing damage to the instrument.

Since the minimum detectable ion current is fixed by the state of the art of d-c amplifier design, it is desirable to obtain as many ions per second as possible at the collector plate. In the ion source the sample pressure and the electron emission from the filament will be as high as is commensurate with reasonable filament life. The slits S_1 and S_2 will be made as wide as possible, resulting in a relatively large angle of divergence of the ion beam and consequently a large focusing error. By referring to Figures 2 and 3 it can be seen that a limitation on slit width and beam divergence is imposed by the relative values of specific masses that are collected and those that are excluded. Examination of equation 10 shows that the separation in space of neighboring masses is greater for low values of specific mass than for high values. For example, the separation between masses 3 and 4 would be

$$\Delta R = R_4(1-3/4)^{1/2} = 0.134 R_4 \tag{11}$$

whereas the separation between masses 39 and 40 would be only

$$\Delta R = R_{40}(1-39/40)^{1/2} = 0.013 R_{40} \tag{12}$$

It is obvious that a mass spectrometer with a small radius of curvature and the widest possible slits is most adaptable to use with low specific masses, and that such an instrument would not be expected to separate neighboring high masses.

The foregoing considerations suggest the choice of helium as the most suitable gas for use as a "probe" or "tracer" gas. Since helium is inert and is rare in the atmosphere, being present to the extent of about one part in 200,000 parts of air at sea level, we would not expect to find an appreciable background signal because of residual helium present in the mass spectrometer tube or leakage of air through any other leaks that may be present. It is also significant that there are not likely to be other gases present which could give an apparent, or spurious, helium signal due to the loss of two or more electrons in the ionization process. For example, the probability of triple charging of carbon-12, or quadruple charging of oxygen-16 in detectable quantities is very small. It is hardly necessary to point out that helium is a gas at all temperatures and pressures to be considered and therefore will not plug leaks or contaminate the system. Neither will it react with any other materials which may be present. The requirements of versatility and certainty of response are seen to be fulfilled by the use of helium as the probe gas. The further requirements of simplicity and portability also are brought closer to realization.

Component Parts of the Leak Detector

In addition to the basic element of the leak detector, the mass spectrometer tube, a number of auxiliary components

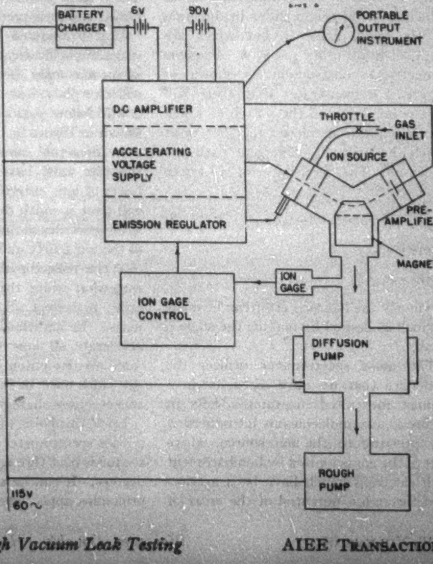

Figure 6. Leak detector block diagram

are required for control purposes. These are shown diagrammatically in the block diagram of Figure 6 and can also be seen in the back view of the instrument shown in Figure 7. The chief parts are as follows:

1. Mass spectrometer tube.
2. Emission regulator.
3. Accelerating voltage supply.
4. D-c amplifier.
5. Ion gauge.
6. Diffusion pump.
7. Rough pump.
8. Throttle valve.

The gas sample from the system under test is introduced to the spectrometer tube through a throttle valve in order that the pressure in the spectrometer tube can be regulated to a suitable value. Some of the gas molecules in the ion source are ionized by bombardment with an electron beam, which is held constant by an electronic emission regulator circuit. The ions are accelerated through a series of plates by an adjustable voltage obtained from the accelerating voltage supply, and are deflected by the field of a permanent magnet with trapezoidal pole pieces which is mounted inside the mass spectrometer tube. The ions which pass through the collector slit are discharged on the collector plate, and the ion current is amplified and measured by the d-c amplifier.

Output current of the amplifier can be read either on a panel instrument or on a portable instrument which is readily placed in view of the operator. While it would be possible to use a well regulated power pack to operate the amplifier, batteries are used for plate and filament voltages in this equipment.

The pressure in the spectrometer tube is measured by a standard ionization gauge with an a-c operated control panel. A protective feature is incorporated in the ion gauge control to automatically turn off the ion gauge and spectrometer tube filaments if the pressure accidentally is allowed to rise above a safe operating limit.

The oil diffusion pump and cold trap, backed by a mechanical rough vacuum pump, will reduce the spectrometer tube pressure to a safe operating value within about an hour after the pumps are started.

The Mass Spectrometer Tube. Figure 8 is a diagram showing the essential parts of the mass spectrometer tube. The permanent magnet produces a field of about 900 gausses, and the tube is designed for a 5-centimeter radius of curvature. Inserting these values in equation 10 we find that helium, of specific mass 4 will be focused on the collector slit with an accelerating voltage of 243 volts. It is also interesting to note that for a constant radius, which is fixed by the slit system, and a constant magnetic field, the product of specific mass and accelerating voltage is a constant, in this case 4 × 243 or 972, so that equation 10 reduces to $M=K/V$. This fact simplifies the problem of identifying other ions that may be collected at different values of accelerating voltage.

Figure 7. Rear view of leak detector showing arrangement of parts

The ion source is somewhat more complicated than is indicated in Figures 2 and 4 due to the addition of an ion lens, consisting of two focusing half-plates which can be separately adjusted to the proper potentials to concentrate the ion beam at slit S_2 and to center the beam on the slit. Referring to the ion lens diagram, Figure 9, if the potential gradient E_b is greater than E_a the ion lens will be converging with a focal length equal approximately to $2V/(E_b-E_a)$, and can be treated in much the same manner as a thin optical lens of the same focal length.[3] Use of such an ion lens increases the intensity of the ion beam, and therefore increases the sensitivity of the leak detector.

The ion source is separated from the remainder of the spectrometer tube by a baffle located at the second source slit. The purpose of the baffle is to introduce additional pumping resistance so that the source pressure will be higher than the tube pressure, and more ions will be formed. The net result is greater sensitivity. This will become apparent if we consider the pressure limitations in the tube.

Any mass spectrometer must be operated under high vacuum conditions for two reasons. First, the average ion must not suffer a collision with gas molecules or other ions while travelling from the ion source to the ion collector, that is, the mean free path must be greater than the length of the mass spectrometer tube. Otherwise the beam will be scattered,

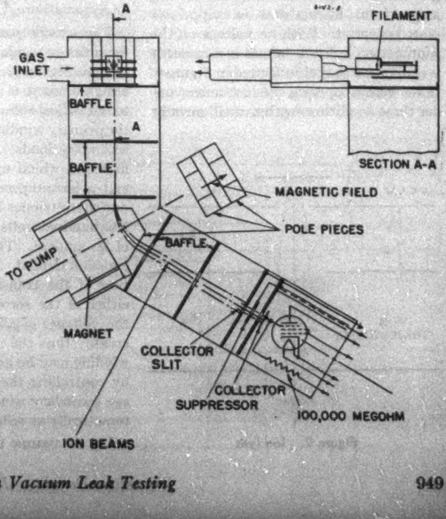

Figure 8. Mass spectrometer tube diagram

with resulting lack of sensitivity, and under certain circumstances a background ion current can appear at the collector. The second reason for the high vacuum is that a tungsten filament is used as a source of electrons, and the life of the filament would be very short at pressures above about 10^{-3} millimeter mercury (1 micron).

Experimental results show that satisfactory filament life under normal operating conditions can be obtained if the source pressure is limited to about 3×10^{-4} millimeter with a filament emission current of five milliamperes. At this pressure, however, the mean free path for air molecules would be less than the length of the path through the spectrometer tube. This condition is obviously unsatisfactory, since more than half of the ions would suffer a collision before reaching the collector. Tests on an actual tube have shown that it is necessary for the mean free path to be more than ten times the length of the spectrometer tube for satisfactory operation, and even in this case an appreciable portion of the ions will suffer collisions.

Use of the baffle allows us to maintain maximum source pressure for highest possible sensitivity and still have a low enough pressure in the major portion of the spectrometer tube to prevent scattering of the ion beam by collisions with air molecules.

The ion collector arrangement indicated in Figures 3 and 4 consists of a slit and a collector plate. Again the actual arrangement used is somewhat more complicated, as shown in the tube diagram Figure 8. A suppressor plate is introduced between the slit and the collector, and a shield is provided to guard the collector from fluctuations in suppressor plate potential. With no voltage on the suppressor plate the actual arrangement is equivalent to that indicated in Figure 4. If we were to operate the instrument under these conditions with a small amount of helium mixed with air at the maximum allowable tube pressure, we would obtain a spectrum in the region of helium as shown by the upper curve of Figure 10. The sloping background curve is due to ions which have been slowed down or deflected by collisions during the transit from the ion source, and which therefore happen to pass through the collector slit. All such ions will have less than normal energy, since each collision results in loss of energy for the high-speed particle. We can erect a potential barrier through which these scattered ions cannot pass by applying a positive potential, essentially equal to the accelerating voltage, on the suppressor plate. We then obtain a spectrum similar to the lower curve of Figure 10, in which it is apparent that the presence of the same amount of helium is much more easily discerned than in the upper curve.

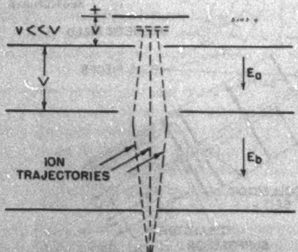

Figure 9. Ion lens

A typical spectrum in the neighborhood of the helium peak for a mixture of 1 part helium in 50,000 parts air is shown in Figure 11.

Production of Ions. Ions are formed in the ionization chamber by bombarding gas atoms or molecules with electrons which have been accelerated through a potential of about 145 volts. The electrons are emitted from an incandescent tungsten filament located behind a slit in the ionization chamber. Since the emission current varies rapidly with changes in voltage, changes in heat conductivity of the surrounding gas, changes in the surface conditions of the filament due to the presence of different gases, and other similar causes it is desirable to hold the emission constant with some kind of an automatic regulator.

The scheme used in the leak detector for emission regulation is shown schematically in Figure 12. A transformer having a high amount of voltage regulation, due to an adjustable resistor in series with the primary winding, is used to supply two secondary loads; the spectrometer tube filament, which operates at a few volts and a few amperes, and a variable load due to the triodes T_3 and T_4 operating at a few hundred volts and a few hundredths of an ampere. The filament is operated emission limited. Due to the high regulation of the transformer, an increase in either of the secondary loads will cause the voltages of all three windings to decrease; thus the voltage on the filament winding may be held at any desired value by controlling the load on the high-voltage secondary winding. This is done automatically as follows:

Let us assume that the filament emission current increases above its normal value. The voltage at the grid of the amplifier tube T_1 will increase due to the increased drop across the emission control resistor; but the cathode voltage of T_1 is held at a constant value by the voltage regulator tube T_2. Therefore T_1 conducts more current, reducing the potential at its plate and lowering the cathode potential of loading tubes T_3 and T_4 with respect to their fixed grid potential. The loading tubes are therefore more conducting and constitute an increased load on the transformer, so that the voltages of all three windings will drop. Since the filament is operating emission-limited and is very sensitive to voltage changes, the emission current will be reduced to a value only very slightly higher than the normal value. In other words, the emission current is regulated at a value determined by the IR drop across the emission control resistor. By changing this resistor, any desired value of emission current can be selected. The circuit is adjusted to the regulating range by means of the adjustable series resistor in the transformer primary. A regulator of this type will hold the emission current constant within less than one-half of one per cent.

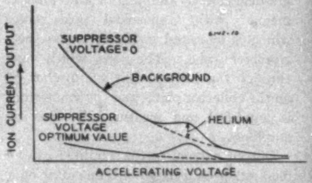

Figure 10. Background characteristics in the region of the helium peak

All voltages shown in Figure 12 are relative values referred to the slit 1 plate, which may be at any value from 0 to 370 volts above ground potential.

The supply voltages for the emission regulator are actually tapped from the ion accelerating voltage supply.

Accelerating Voltage Supply. The ion accelerating and focusing potentials are obtained from a conventional full-wave rectifier with voltage regulator tubes for d-c output voltage regulation and a simple voltage divider circuit shown schematically in Figure 13. The accelerating voltage (between slit 1 and ground) is adjusted with potentiometer R_1, which actually consists of a coarse and a fine adjustment. This adjustment is used to bring helium ions through the collector slit.

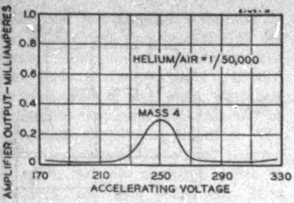

Figure 11. Typical helium spectrum from leak detector

Ion Current Measurement. Attainment of maximum over-all sensitivity implies the use of the most sensitive available method of ion current detection; but the detector must also meet the important requirements of ruggedness, simplicity of operation, and portability. A d-c amplifier with 100-per-cent feedback meets these requirements satisfactorily, although it is somewhat less sensitive than certain other methods, such as an electrometer tube and sensitive galvanometer. The amplifier, when used with an input impedance of 10^{11} ohms and an output instrument rated 30 millivolts full scale, is capable of measuring ion currents of the order of 10^{-14} ampere.

The amplifier consists of an acorn tube connected as a space-charge tetrode followed by two stages of voltage gain and a cathode follower output tube. The output appears on an ordinary panel-type milliammeter rated one milliampere full scale, and having a resistance of 30 ohms. The amplifier circuit is shown schematically in Figure 14A. Such an amplifier possesses the valuable characteristic that the output voltage is equal to the input voltage, but reversed in sign. This can be shown to be the case by consideration of Figure 14B. We will define input voltage as the voltage drop across the input resistor R_i due to the flow of ion current I, where the input resistance is very high (of the order of 100,000 megohms) and the ion current is very small (less than one micromicroampere), so that so far as the ion current is concerned the negative side of the input resistor is grounded through the low impedance millivoltmeter circuit. The output voltage is defined as the voltage from B to ground, which is equal to the voltage at C minus the output battery voltage E_B. The output voltage can be balanced to zero when there is no ion current flowing by adjusting the plate potentiometer in the first stage of the amplifier until the voltage at C is equal to the battery voltage.

An ion current I flowing through the input resistor will produce an input voltage V_i, and point A will go positive by an amount ΔV which for the moment we will assume to be any value not greater than V_i. By reference to Figure 14A we can see that since the last stage is a cathode follower, the amplifier will behave as if it had an odd number of stages, and there will be a phase reversal. Therefore the voltage at C will decrease by an amount $\mu\Delta V$, where μ is the over-all voltage gain of the amplifier, provided by the second and third stages. The voltage at C changes from the initial value E_B to a new value $(E_B-\mu\Delta V)$, and the voltage at B, which is controlled by the stiff output circuit, changes from zero to a new value

$$V_o=-\mu\Delta V \tag{13}$$

Applying Kirchhoff's law to the circuit from ground to B to A and back to ground, we obtain

$$-\mu\Delta V+IR_i-\Delta V=0$$

or

$$IR_i=(1+\mu)\Delta V \tag{14}$$

If the gain μ is much greater than unity this becomes

$$IR_i=\mu\Delta V \tag{15}$$

But IR_i is defined as the input voltage, and $\mu\Delta V$ is the negative output voltage. Therefore we write

$$V_o=-V_i \tag{16}$$

It is apparent that ΔV is very small in comparison with the input voltage, which means that point A will always remain at essentially ground potential, so that the apparent input resistance of the amplifier taken from A to ground will be less than the actual resistance by the factor $1/\mu$. The time constant of the input circuit is reduced accordingly by the same factor, which is an important feature in improving the over-all response of the leak detector. It is also evident that the output voltage cannot exceed the battery voltage E_B, since the voltage at C cannot have a negative value. The output instrument therefore is protected against excessively large input signals.

The first stage of the amplifier is mounted inside the mass spectrometer tube as shown in Figure 8 in order to keep the high resistance and the grid connections dry to minimize leakage currents, to keep the leads as short as possible from collector plate to the grid of the first tube, and to provide electrostatic shielding for the input stage. The remainder of the amplifier is mounted in the top part of the leak detector control panel, as indicated in Figure 6.

Ionization Gauge Tube and Control.* The pressure in the high vacuum system is measured with a standard ionization gauge (Distillation Products type VG1-A) and an electronic control circuit. The gauge tube consists of a hairpin filament surrounded by a grid which is wound in a double helix with wide turn spacing, and a plate or collector consisting of a platinized layer on the pyrex glass envelope of the tube. Both ends of the grid are brought out, so that the grid can be outgassed by passing current through it.

Electrons emitted by the incandescent filament are accelerated through the grid, which operates at positive potential with respect to the filament. Since the collector is at a negative potential with respect to the filament, the electrons will return to the grid and be collected. Some of the electrons will collide with gas molecules, causing ionization in the space between the grid and the collector. The ions are drawn to the collector and discharged. With constant emission current and constant grid potential, the ion current is a linear function of pressure for any gas, although the calibration will vary somewhat for different gases.

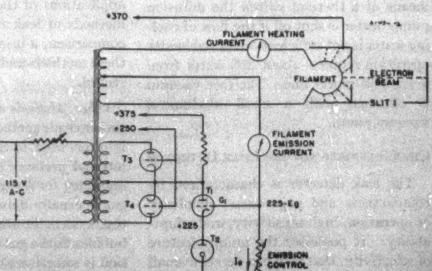

Figure 12. Emission regulator

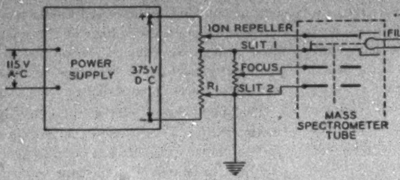

Figure 13. Accelerating voltage supply

The ion gauge control is an a-c operated unit which performs four functions:

1. Voltage supply for ionization gauge tube.
2. Emission regulation and emission current measurement.
3. Ion current measurement.
4. Vacuum relay to operate if the pressure in system exceeds a predetermined value.

The direct voltage supply is a conventional full-wave rectifier, with an output of several milliamperes at 400 volts.

The emission regulator circuit is essentially the same as shown in Figure 12.

The ion current is measured with a simple vacuum tube microammeter circuit shown in Figure 15. A full-scale deflection on the highest sensitivity corresponds to one-half-microampere input.

The vacuum relay circuit is also shown schematically in Figure 14. The relay is used to turn off the ion gauge and spectrometer tube filaments in the event that the pressure accidentally rises above a safe operating point. Since samples are continually connected to and removed from the leak detector in normal operation, it is inevitable that the wrong valve will eventually be left open. This relay has saved many filaments.

Pumping System. The vacuum in the leak detector is maintained by an oil diffusion pump with a cylindrical dry ice trap. The speed of this combination is about 30 liters per second. The diffusion pump requires about 3 cubic centimeters per second of tap water for cooling. By means of a thermal switch the diffusion pump heater is shut off if the flow of cooling water is not sufficient. Normal heater operation requires about 350 watts from a 110-volt power line. The fore vacuum is obtained with a small mechanical vacuum pump.

CHARACTERISTICS OF THE LEAK DETECTOR

The leak detector is characterized by compactness and ruggedness, simplicity of operation, high sensitivity, and adaptability. It possesses the unique feature of selectivity, that is, it can detect a small leak in the presence of a large leak, or it can detect a leak in the presence of thousands of other leaks of the same size. It is therefore a very useful method for testing very large vacuum systems which may have a multitude of leaks.

The sensitivity of the leak detector is such that it will detect the helium in normal air (1 part helium in 200,000 parts air). Under these conditions the partial pressure of helium in the leak detector is only 5×10^{-11} millimeter Hg measured at the ion gauge tube. The quantity of helium flowing through the spectrometer tube is about 1.5×10^{-9} liter-millimeter per second, which is roughly equivalent to a flow of 1 cubic centimeter in 16 years at standard temperature and pressure.

Since the response of the leak detector depends only upon the presence of helium in sufficient quantities, and not upon the purging of the system by the probe gas, the response of the instrument when used on small systems is almost instantaneous for practical purposes. The signal will disappear as fast as the helium is pumped from the system, which is usually a matter of a few seconds but depends upon the size and pumping characteristics of the sample system. Test methods will be discussed in the following section on applications.

APPLICATIONS OF THE MASS SPECTROMETER LEAK DETECTOR

The full degree of flexibility of the mass spectrometer as a leak detector cannot be visualized without consideration of its applications. Its advantages are more apparent when compared with the limited applications of the other commonly used methods of leak testing. To aid in this comparison, a brief discussion of some of these methods and their limitations is presented.

Other Methods of Leak Testing. A rather crude method of leak testing which is often used is that of building up the internal pressure of the test system and searching for escaping gas. The escaping gas is usually detected by immersing the test system in water and observing the bubbles as the gas escapes. This observation is sometimes made more sensitive by adding chemicals to the water to reduce the surface tension.

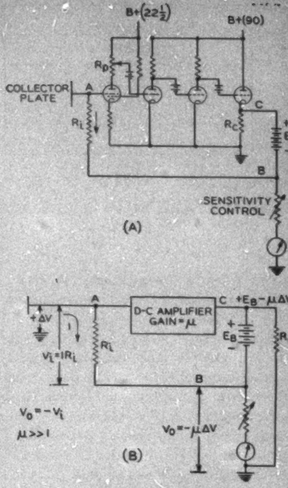

Figure 14. D-c amplifier

If certain gases are used to pressurize the system, the escaping gases may be observed by noting a change in color of a gas flame used as a probe. This method is restricted to large leaks in systems which can withstand internal pressures of several atmospheres.

Several methods of leak testing are in use which depend upon the detection of gases leaking into an evacuated system. The means of detecting these leaks are as varied as the properties of gases themselves. Mention need be made of only the few which have been found most useful.

One method of detecting such leaks makes use of the color of the glow discharge of different gases. If the system under evacuation has an air leak, a glow discharge induced by a high-frequency high-voltage supply will have the characteristic pink color which indicates the presence of nitrogen in the system. The leak may be found by observing the change in color of the discharge as the leak is covered by a gas or liquid which has a glow of different color. Examples are the grayish glow of oxygen and the blue glow of ether. This method is applicable primarily to glass systems in which the glow discharge can be easily initiated and observed. It is limited to pressures that will maintain a glow discharge, and therefore cannot be applied to leaks large enough to give pressures greater than a few millimeters of mercury or to leaks in a system where the pressure is lower than a few microns. If multiple leaks exist the discharge is masked by air leakage and the sensitivity is decreased. Due to its low sensitivity, the glow discharge method cannot be applied to the location of small leaks. However, indications that small leaks exist may be obtained by the following procedure. The system is evacuated to a good vacuum, in which the glow discharge cannot be initiated, and then sealed off. If a later check shows a pink discharge, an air leak is assumed to exist. This may be used to indicate very small leaks, but often requires storage of the evacuated system for several weeks. Also, observed results sometimes are difficult to interpret.

A leak in a system under evacuation sometimes may be located by covering the leak with water or other liquid having a vapor pressure of less than one atmosphere and observing the pressure drop in the system. This method is limited by the accuracy with which a small pressure change can be detected and does not apply at all to small leaks. Indicated results are often doubtful or misleading.

If the system being evacuated has a pressure greater than 0.1 micron, the heat conductivity of the gas is sufficient to produce a detectable cooling of a heated filament. Since the heat conductivities of hydrogen and helium are about six times that of air, large leaks may be found by observing the change in resistance of a heated filament when one of these gases is used for probing. In addition to being limited to large leaks and certain pressure conditions, this method requires differentiating between heat conductivity changes due to the probing gas and heat conductivity changes due to changes in pressure.

The temperature-limited emission from a tungsten filament depends upon the condition of the surface of the tungsten. In a good vacuum, a small increase in the concentration of oxygen will produce an observable decrease in this emission. Therefore, leaks may be found by observing the decrease in emission from a constant temperature filament while using oxygen as a probing gas.[3] Under favorable conditions this method of leak testing is very sensitive. For this reason it has found wide application to laboratory vacuum systems. Its disadvantages are that an extremely good vacuum is required, the filament must be initially clean, and the quantity of oxygen and water vapor evolved from the walls of the vacuum system must be very small. Although this method of leak testing has high sensitivity, it is not as sensitive as the mass spectrometer method.

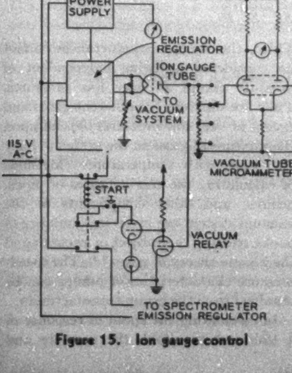

Figure 15. Ion gauge control

Another method often used is that of making repeated checks on the rate of pressure rise in a sealed-off system after successively painting small areas with compounds which will plug the leaks. This method is time consuming and the results are not conclusive.

Mass Spectrometer Leak Detector. The characteristics of the mass spectrometer as a leak detector already have been discussed. The absence of the limitations of the other methods make it applicable to many new fields of leak testing. In fields where other methods have been used, it provides an easier, faster, and more reliable means of leak testing.

Leak Test Vacuum Systems. The use of the mass spectrometer leak detector requires a vacuum manifold and an auxiliary pumping system. The design of the vacuum manifold and pumping system is determined by the leak test requirements. Some of the factors which influence the design are

1. Size and shape of test objects.
2. Number of objects to be tested.
3. Expected evolution of gas from the test objects.
4. Sensitivity required.
5. Time allotted to leak testing.

Several general types which meet the majority of applications will be discussed.

Figure 16 is a block diagram showing the simplest form of leak test manifold. In using a manifold of this type, the procedure is to first evacuate the test objects by opening the exhaust valve and the isolating valves. In well-designed manifolds of moderate size, this rough evacuation takes less than 30 seconds. The throttle valve is then opened as far as possible without exceeding the permissible pressure in the spectrometer tube, and the system is ready for leak testing. Since the response to a leak is proportional to the rate at which helium passes through the spectrometer tube, the greatest sensitivity is obtained by opening the throttle valve and closing the exhaust valve to maintain the maximum permissible pressure in the spectrometer tube. The ultimate sensitivity is obtained when the throttle valve is wide open and the exhaust valve is closed entirely. All of the gas from the test manifold and test objects then passes through the spectrometer tube.

The two obstacles to obtaining this ultimate sensitivity are large leaks and the evolution of gas from the walls of the test objects and test manifold. In the first case, the ultimate sensitivity is not required and the leaky test objects can be located and removed from the system by closing their isolating valves. In the second case, a liquid air or dry ice trap in the line between the test manifold and the spectrometer tube will collect the condensable gases, but not helium, and thus allow a larger portion of the helium to be passed through the spectrometer tube. The amount of gas evolved is reduced by using a smaller test manifold. With a small manifold the ultimate sensitivity can be obtained without the cold trap mentioned above. This type of manifold is thus seen to be particularly applicable to high sensitivity requirements on vacuum systems where the amount of evolved gas is not great.

In applications where the maximum sensitivity is not required, but the time allotted to leak testing is short, a larger manifold having a more elaborate pumping system is better adapted. A system to meet these requirements is shown in Figure 17. This system is really two manifolds served by a single set of pumps and a single leak detector. The method of operation is quite similar to that of the simple manifold. Assume that initially

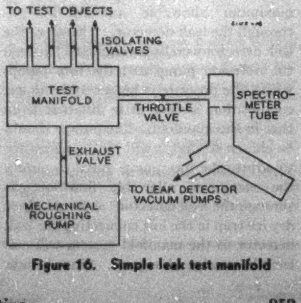

Figure 16. Simple leak test manifold

both roughing valves, both exhaust valves and the throttle valve are closed. Test objects are placed on manifold number 1 and evacuated to a rough vacuum by opening the isolating valves and the number 1 roughing valve. When the pressure in the manifold is reduced to a few microns, roughing valve number 1 is closed and exhaust valve number 1 is opened. The throttle valve then is adjusted to give the maximum allowable pressure in the spectrometer tube and manifold number 1 is ready for leak testing. While leak testing continues on manifold number 1, test objects are placed on manifold number 2 and evacuated to a rough vacuum by opening the isolating valves and the number 2 roughing valve. Thus the rate of leak testing can be more than double that obtained with only one of the manifolds.

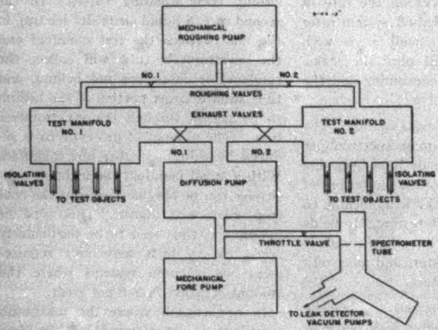

Figure 17. Dual leak test manifold for high speed testing

On systems of the type just described, the manifolds usually are made large so that a large number of test objects can be accommodated. With these larger manifolds it is advantageous to use a diffusion pump to increase the speed of response to a leak. If the leak detector is connected between the manifold and the diffusion pump, the pressure in the manifold may be low enough to prevent passing the maximum allowable amount of gas through the leak detector. Therefore the leak detector usually is connected between the diffusion pump and the fore pump, where the pressure is higher and the gas composition is very little different from that in the manifold. The pumps should be chosen so that it will not be necessary to adjust the fore pump valve to obtain the maximum allowable flow of gas through the leak detector. A liquid air or dry ice trap in the line connecting the leak detector to the manifold system will collect the condensable gases. Since helium will not be condensed. the gases passing into the leak detector will contain a greater proportion of helium and the system will have greater sensitivity.

The vacuum system for the leak testing of large tanks, climatized chambers, and so forth, does not differ greatly from those shown in Figures 16 and 17 if the manifolds are replaced by the tank to be tested. The time of response to a leak increases with increase in volume and decreases with increase in speed of the pumping system. For this reason, it is desirable to use high speed vacuum pumps when testing large systems. In designing the pumping system for these applications, proper consideration should be given to the number of objects to be tested.

A method of leak testing which is well adapted to certain production line manufacture, requires the evacuation of the test objects to a rough vacuum and sealing them off. The evacuated test objects are passed into a chamber containing a high concentration of helium and left there for a specified time. They then are removed and each one checked on the leak detector to determine if helium has leaked into its vacuum. The sensitivity of this method increases with the length of time the test objects are left in the helium chamber. After leaky test objects have been found by this method, the leaks can be located accurately by probing with a helium jet. This method may be extended to locating test objects with very minute leaks, if they are left in the helium chamber for a long time. However, the point of leakage cannot be located by probing if the leakage rate is less than 2×10^{-9} cubic centimeter of air per second (measured at 0 degrees centigrade and 760 millimeters pressure).

Other methods of using the leak detector for leak testing have been tried and still others are possible, but their applications seem too limited to merit description at present.

Analogy of Vacuum Systems to Electric Network. In the design of vacuum manifolds for leak testing, two considerations are predominant. The response to a leak should be fast and the percentage of helium from the leak passing through the leak detector should be high. These two considerations are not always independent and a compromise based on the particular application must be made. In the analysis of the vacuum system and its effect upon leak testing, it is often advantageous to think of the vacuum system in terms of its electric network analogue. A list of corresponding terms is given.

Electrical Entity	Vacuum Equivalent
Voltage, V (volts)	Pressure, P (microns)
Quantity of electricity, Q (coulombs)	Quantity of gas, PV (micron-liters)
Current, $I=dQ/dt$ (amperes)	Rate of gas flow, $Q=d(PV)/dt^2$ (micron-liters per second)
Conductance, G (mhos)	Conductance, F, S (liters per second)
Capacitance, C (farads)	Volume, V (liters)
Inductance, L (henrys)	Inertia of gas (negligible in vacuum)

In using this analogy there are several things which should be noted. First, the symbols generally used may occur in both systems of units, but do not necessarily represent corresponding entities. Second, in vacuum terminology, speed and conductance are not separate entities and are often used interchangeably. However, speed generally is used to denote the conductance of a pump or the net conductance of a pump and connected system. For example, a pump of speed S, connected through a pipe of conductance F, has a resultant speed of S_R given by

$$\frac{1}{S_R}=\frac{1}{S}+\frac{1}{F}$$

Third, the vacuum conductances cannot be considered as being independent of pressure except at very low pressures. They are different for different gases and except at very low pressures depend upon the mixture of gases in the system. They also vary with temperature. Methods of calculating the conductances of pipes, orifices, and other components of the vacuum system are given in a number of books treating the subject of vacuum and the kinetic theory of gases.[4,5] The speed-pressure characteristics of pumps can be obtained from the pump manufacturers.

In calculating the speed of response of a leak test manifold, the pressure and temperature may be considered as constant during the leak testing. The conductances may be calculated for the pressure and temperature considered and treated as constants. Since the leak detector signal is proportional to the partial pressure of helium in the manifold, the rate at which the helium pressure builds up and decays may be taken as the speed of response of the manifold. Consider the simple system shown in Figure 18A, consisting of a pump of speed S liters per second connected to a volume of V liters in which the pressure P microns, remains uniform. Assume that this system has a constant leak of Q micron liters per second. The electrical analogue is shown in Figure 18B. It is a capacitor of C farads with a discharge resistor of R ohms and a constant charging current of I amperes. Applying the charging current when the voltage is zero, the voltage build-up is given by

$$V=IR\left(1-\epsilon^{-\frac{1}{RC}t}\right)$$

When the charging current is removed from the fully charged capacitor the voltage decay is given by

$$V=IR\epsilon^{-\frac{1}{RC}t}$$

In the vacuum system the application of the helium jet to the leak and its removal give the corresponding equations

$$P_{He}=\frac{Q_{He}}{S}\left(1-\epsilon^{-\frac{S}{V}t}\right)$$

and

$$P_{He}=\frac{Q_{He}}{S}\epsilon^{-\frac{S}{V}t}$$

where P_{He} and Q_{He} refer, respectively, to the partial pressure and rate of leakage of helium. When considering long pipes in vacuum systems, the pressure is not uniform throughout the length of the pipe. The electrical analogue is a d-c transmission line with distributed resistance and capacitance.

Fields of Leak Detector Application

It is impossible at this time to compile a comprehensive list of the different fields in which the leak detector will be applicable. However, a list of those fields in which its application has been studied will serve to show the scope.

Vacuum valves
Vacuum bellows
Vacuum seals
Vacuum chambers of all sizes
Vacuum tubes
Power rectifiers
Hermetically sealed instruments
Vacuum pumps
Pressure containers
Refrigerators
Pressure gauges
Pressure instruments
Vacuum instruments
Vacuum welds
Laboratory vacuum systems
Cyclotrons
High voltage accelerating tubes

The leak detector may be used to reject defective parts or it may be used to accurately locate leaks so that parts may be repaired. It finds additional application as an aid in selecting production methods and in locating faulty processes. These applications are well illustrated by an actual example.

In the manufacture of a large power tube, rejections due to leaks were excessive. Investigation by older methods of leak testing were slow and results were inconclusive. The application of the leak detector provided a fast and reliable method of investigation. Practically all of the leaks were found to occur in one of the small subassembly parts. Pre-leak testing of this subassembly was initiated and rejections were made on the relatively inexpensive subassembly instead of on the completed tube. Accurate locating of leaks in the subassembly led to improved methods in its manufacture so that very few rejections occurred. The over-all result was increased production, lower manufacturing cost, and improved quality of the product.

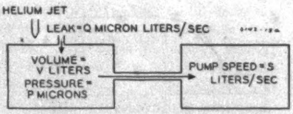

Figure 18A. Simple vacuum system

Special Leak Testing Considerations

On large manifold systems where rapid testing of a large number of objects is required, it has been found that testing time can be reduced by the use of a testing hood to surround the entire test object with an atmosphere of helium. This eliminates probing time and gives a response proportional to the total leakage if several leaks exist. The possibility of missing leaks by careless probing is also removed. After locating leaky test objects the hood may be removed and the leaks located accurately by probing with a small jet of helium. A valuable adjunct to the testing hood is a thermal conductivity gauge calibrated to read directly the concentration of helium in the hood.

For most probe testing, probes made from 4-inch metal capillary tube of 5- to 20-mil diameter with one pound pressure of helium have been satisfactory. However, when testing in confined spaces where the helium cannot readily diffuse away, the helium concentration may become sufficient to give a reading when the probe is some distance away from the leak. In this case a special probe may be made which will withdraw the excess of helium. Spurious signals may sometimes be caused by air currents carrying helium from the probe to a large leak some distance away. These are recognized easily by their erratic nature and may be eliminated by locating the leak causing the trouble and isolating it from the system.

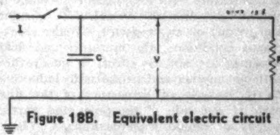

Figure 18B. Equivalent electric circuit

Very small leaks are sometimes temporarily plugged by dust particles or water vapor and may escape detection at the time of leak testing. Heating seems to be the most effective way of reopening these leaks.

Results obtained by the leak detector are repetitive and the leak detector can be used to measure rate of leakage by comparison with a calibrated leak. Also, the sensitivity of a particular leak test system may be obtained by observing the magnitude of the signal from a calibrated leak attached to the system.

Other applications of leak detectors are indicated in which helium may be used as a tracer gas. These would include investigation of flow in ventilating systems, velocity and distribution of flow in piping systems, turbulence, and other similar problems.

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Current and Torque of D-C Machines on Short Circuit

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Synopsis: This paper presents formulas, in terms of easily evaluated coefficients, for the armature current, field current, and torque of d-c machines on short circuit. The method of the paper makes it unnecessary to calculate the mutual inductance between the armature and field windings. Heretofore, this mutual inductance (between the two windings which normally are in quadrature with each other) has made it exceedingly difficult to calculate current and torque on short circuit. Under short-circuit conditions, the armature and field windings are not, in effect, in quadrature with one another and are mutually inductive partly because of saturation of the flux path at one side of the poles and partly because of commutation phenomena. These circumstances create difficulties uniquely overcome in the paper. A basic feature of the paper is the expression for the relation between flux, armature amperes, and field amperes. Calculations cannot be made without establishing this relationship (plotted for a specific case in Figure 9) and a simple expression for it has not been presented before to the author's knowledge. Another important feature, also believed to be new, is the development of an expression for the resistance to current flow created by the reactance voltage of the alternating current in the armature winding. A third feature is a formula for estimating the time constant of the armature circuit using an average coefficient of inductance. The results are applied to several machines and the calculated performance compared with test data. In most cases, particularly with large machines having bar windings, approximate calculations can be made without need of manufacturer's data.

PROTECTION must be provided for d-c machines and d-c distribution circuits against the large currents which flow on short circuit. Otherwise, machines, switchgear, instruments, and bus structures will be damaged by overheating, flashing, and burning. When machines are connected in parallel to a common bus, as in the case of continuous rolling mills or shipboard distribution circuits, the electromagnetic forces between current-carrying parts of the equipment may be astonishingly large. Also, it has happened that the sudden high torque of the machines has twisted shaft keys breaking coupling hubs. In applying circuit breakers, it has been necessary to calculate the current and voltage at the time of interruption for various types of breakers of different operating speeds. It has been equally important to build all machines strongly enough to withstand mechanical stresses and, therefore, to calculate maximum torque and forces on the windings and their supports at maximum current. These calculations have been uncertain and stresses assumed beyond those which actually exist, in spite of the fact that size, weight, and cost are sometimes determining factors in the selection of machinery and its successful application.

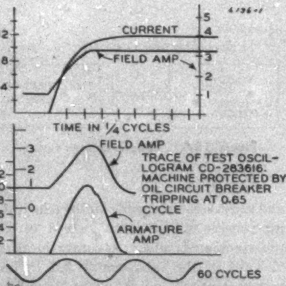

Figure 1. Currents on short circuit of d-c generator

Type *MCF*, 8 poles, 1,500 kw, 514 rpm, 1,500 volts

$i_a=3.3+9.2\epsilon^{-1.8t}-12.5\epsilon^{-206t}$

$i_f=1+(3.3-0.7)(\epsilon^{-1.8t}-\epsilon^{-206t})$

$\rho=0.17$, $n_f=0.15$, $R=0.040$,

$K_1=6.4$, $K_2=0.2$, $\phi_0=1$

Armature axis shift $=1\frac{1}{4}b$

It is helpful to recognize that at the moment when the current reaches its maximum the internal voltage of the average d-c machine is a high fraction of the voltage before short circuit. As there is no large immediate collapse of internal flux and voltage, the maximum current is determined chiefly by the resistance of the windings and by the a-c reactance voltage in the armature winding which causes an equivalent large d-c brush contact voltage. It is the purpose of this paper to emphasize and establish these phenomena, and to develop formulas for calculating the current and torque.

The calculation of peak current is given particular attention because the sustained current is not reached until induced current in the exciting winding has died out. Exciting windings of d-c machines are notoriously slow, and the induced field current takes more time to disappear than is needed for operating even slow acting breakers. Moreover, large machines will flash over if the current is not interrupted quickly, and damage will be done to circuits, apparatus, and instruments. Therefore, circuit breakers are used nearly always in the armature circuit, capable of opening peak short-circuit current. Consequently, the sustained current is not often of practical importance. However, for completeness, calculation of the sustained current is included in the paper.

The results of the paper and general experience show that the maximum current on "dead" short circuit is 7 to 15 times rated current and maximum torque 5 to 12 times normal torque. In general, the higher values are encountered with 600- to 1,500-volt machines, and the lower values with 125- and 250-volt machines. Slow speed machines generally give lower values than high speed machines. Machines with pole face windings have higher values than those without. Also, values are higher if the shape of the armature is a cylinder of short length rather than long.

That the current is not much higher is surprising in view of the high internal electromotive force immediately after short circuit and the low normal resistance of the windings. Ordinarily, rated voltage divided by the normal resistance gives 25 to 50 times rated current. This indicates the importance of the reactance voltage produced by the alternating current in the armature winding, in limiting the current, a major feature treated in the paper.

Evidence that the electromotive force does not collapse immediately is obtained from observations of the behavior of the field current on many short-circuit tests. Many of these have been made over the years in connection with tests of circuit breakers using d-c machines for power. Usually the field current rises to not more

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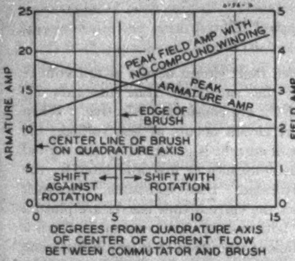

Figure 2. Effect of brush arcing (shifting axis of armature magnetomotive force) on armature and field currents

Type *MCF,* 8 poles, 1,500 kw, 514 rpm, 1,500 volts

than three times its value before short circuit. Also, it is well known that except for small or special machines the leakage flux linkages of the exciting winding, producing no electromotive force in the armature, are a small part (15 per cent or less) of those producing electromotive force. Therefore, to illustrate the evidence with numbers, the total linkages before short circuit may be taken as 115 per cent. Then, assuming the field current rises to three times the initial value after short circuit, the leakage flux interlinkages increase correspondingly to 45 per cent, and the flux which produces electromotive force, according to the constant flux linkage theorem, is 115 per cent minus 45 per cent, or 70 per cent. Thus the electromotive force after short circuit is 70 per cent of the initial electromotive force, until the unsupported field current decays. This decay occurs after the maximum current has been reached. In fact, the field and line currents rise to a maximum at the same instant.

This may make it appear to the practical minded person that all one needs to do to calculate the maximum current is to take 70 per cent of the rated voltage and divide by the resistance of the windings. For a rough approximation, the calculation is as simple as this provided one allows for the a-c reactance voltage in the armature. Then the maximum torque may be estimated in per cent of rated torque simply by taking the maximum current in per cent of rated current and taking 70 per cent of it.

G. E. Frost recently presented a paper[1] in which it was shown that the armature and excitation circuits are coupled together magnetically with a coefficient of mutual inductance *M*. In normal operation the armature magnetomotive force and flux are in the cross or quadrature (interpole) axis with respect to the excitation (field) magnetomotive force and flux in the direct or field pole axis. Therefore, the armature and excitation circuits ordinarily are not coupled if there is no compound winding. *M* becomes a real quantity only when the armature magnetomotive force becomes large with overload current, causing saturation of the armature teeth at one side of the main pole with an appreciable shift in the effective position of the main pole axis of the machine. Furthermore, when the commutating poles become saturated and their flux is no longer sufficient to hold the current collection from the commutator central under the brushes, the axis of armature magnetomotive force shifts to reduce its angle with the effective main pole axis still more. The factors which Frost used account for these phenomena, but it is difficult to evaluate them numerically by calculation or to measure them by test. The present paper builds on this essential basic theory. In doing so, machines with and without pole face windings are kept in mind, making the method applicable to all types. Pole face windings hold the armature and field circuits in a more nearly true quadrature relationship, making it difficult to account for test results by the theory of mutual coupling between these circuits. The viewpoint from machines with pole face windings leads directly to recognition of the major factors which determine the maximum (peak) current.

Assumptions

1. There are no shunts across pole face or commutating pole windings.

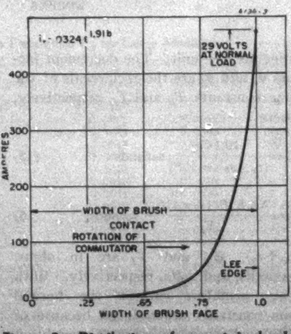

Figure 3. Distribution of current in brush face of d-c generator

Type *MCF,* 8 poles, 1,500 kw, 514 rpm, 1,500 volts

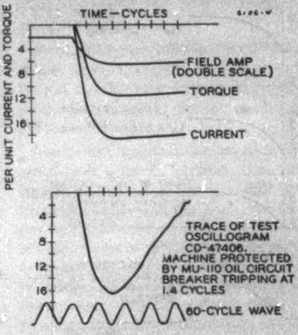

Figure 4. Currents and torque on short circuit of d-c generator

Type *MCF,* 6 poles, 750 kw, 600 rpm, 1,500 volts

$\rho = 0.215$, $n_f = 0.15$, $R = 0.033$
$K_1 = 6.5$, $K_2 = 0.24$, $\phi_0 = 1$
$i_a = 8.2 + 11.4\epsilon^{-1.8t} - 19.6\epsilon^{-143t}$
$i_f = 1 + 2.3(\epsilon^{-1.8t} - \epsilon^{-143t})$
$T = 0.033 i_a{}^2$

2. Saturation other than in armature teeth and pole face teeth is neglected.

3. Leakage flux interlinkages of the field winding are proportional to field amperes.

4. Center of current flowing between brushes and commutator is at lee edge of the brushes. However, provision is made for modification of this assumption in exceptional cases.

5. Speed of the machine is maintained.

Per-Unit Quantities

Per-unit quantities, used throughout the paper, have the following base or unit values:

V = unit voltage = rated voltage
I_{a1} = unit armature amperes = maximum continuous rated amperes
I_{f1} = unit field amperes = field amperes for rated voltage at rated base speed, no load
M_{f1} = unit magnetomotive force ampere turns = field magnetomotive force at unit field amperes
ϕ_1 = unit flux = field flux corresponding to unit field amperes (excluding field leakage flux)

Formulas for Current and Torque

Complete formulas for armature current, field current, and torque, each as a function of time, are derived in Appendix I. The method used consists of the following steps:

1. Develop the general equation for the relation between exciting flux, armature amperes, and field amperes (equation 23).

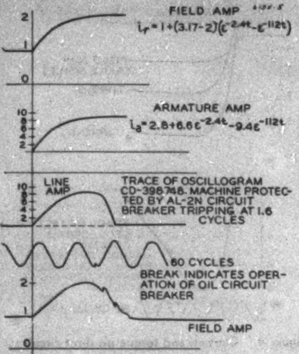

Figure 5. Currents on short circuit of d-c generator

Type *MPC*, 6 poles, 300 kw, 1,200 rpm, 120/240 volts

$\rho = 0.186$, $R = 0.070$, $n_f = 0.107$

$K_1 = 2.06$, $K_2 = 0.214$, $\phi_o = 1$

Line and circuit breaker inductance = 0.26 per unit

2. Equate the field linkages immediately after short circuit (at the instant of peak current) to those immediately before short circuit in accordance with the constant linkage theorem (equation 27).

3. Write the equation summing the voltages in the armature circuit (equation 28).

4. Solve the equations simultaneously for the currents.

5. Combine the currents with their respective decrement factors governing their rise and decay. These decrement factors are the reciprocals of the time constants for the armature and field circuits derived in Appendix II.

The results follow.

For armature current:

$$i_a = I_a - I_a'\epsilon^{-\sigma_a t} + (I_a' - I_a)\epsilon^{-\sigma_f t} \tag{1}$$

For field current:

$$i_f = I_f - (I_f - I_f')(\epsilon^{-\sigma_f t} - \epsilon^{-\sigma_a t}) \tag{2}$$

For torque:

$$T = \frac{Rn_1}{n} i_a^2 \tag{3}$$

where

$$I_a' = -\frac{C_1}{2} + \sqrt{\frac{C_1^2 + 4C_2}{4}} \tag{4}$$

$$I_f' = \frac{(1+n_f)\phi_o - I_a'R}{n_f} \tag{5}$$

$$I_a = -\frac{C_3}{2} + \sqrt{\frac{C_3^2 + 4C_4}{4}} \tag{6}$$

$$C_1 = \frac{K_1RB + 2K_1n_fbP - BK_2(1+n_f)\phi_o}{Rn_fB + K_2RB + 2K_2n_fbP} \tag{7}$$

$$C_2 = \frac{BK_1(1+n_f)\phi_o}{Rn_fB + K_2RB + 2K_2n_fbP} \tag{8}$$

$$C_3 = \frac{2K_1bP - BK_2}{RB + 2K_2bP} \tag{9}$$

$$C_4 = \frac{K_1B}{RB + 2K_2bP} \tag{10}$$

$$K_1 = \frac{\phi_t p_a p_p}{p_b(p_a M_{a1} + p_p M_{b1})} \tag{11}$$

(When there is no pole face winding, $p_b = p_a$ and $M_{b1} = 0$)

$$K_2 = \rho K_t = \frac{gK_\lambda C_s C_d}{0.85(g+d_s)} K_t \tag{12}$$

$$R = r_w + r_b = r_w + \frac{E_B b + V_c}{V} \tag{13}$$

When E_B is not known, it may be estimated at rated load (expressed in kilowatts) and rated speed by

$$E_B = \frac{C_r \times \text{kilowatts}}{d_a} \tag{14}$$

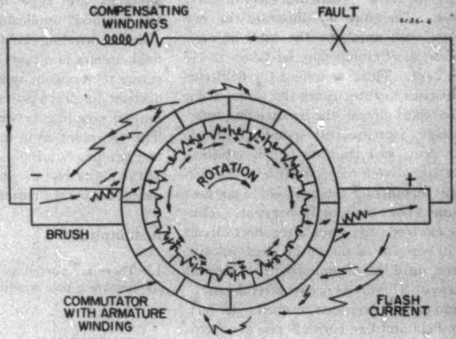

Figure 6. Current between commutator and brush—diagram showing that flash current flows opposite in direction to brush current

Definitions of symbols are given in Appendix VI. All of the foregoing quantities are per unit. The decrement factors σ_a and σ_f are the reciprocals of the time constants T_a and T_f, respectively, where

$$T_a = \frac{19.1C_x}{(r_w + r_b)(n_1 \times P)} \text{ seconds} \tag{15}$$

$$T_f = \frac{N\phi_1 K_t P(1+n_f)10^{-8}}{V_f} \text{ seconds} \tag{16}$$

$(r_w + r_b)$, ϕ_1, and V_f are in ohms, maxwells, and volts, respectively. With heavy armature current, the normal time constants are reduced because of additional saturation. For this condition the "saturated" time constants are

$$T_{as} = T_a \frac{\rho + n_f}{1 + n_f} \tag{17}$$

$$T_{fs} = T_f \frac{\rho + n_f}{1 + n_f} \tag{18}$$

When short circuit occurs at full voltage only the "saturated" time constants T_{as} and T_{fs} are considered effective. When short circuit occurs at less voltage, the rise of current is governed by the normal time constants at first and later by the "saturated" time constants. In such cases the current-time curves are plotted for both sets of time constants. Then a judicious transition is drawn in from one to the other. For general calculations, only the "saturated" constants should be used.

Comparison of Calculated and Test Results

Calculated results are presented for comparison with test oscillograms for three machines on which short-circuit tests have been made. In making the tests, the machines are protected with circuit breakers interrupting the current almost as quickly as the peak is reached.

In Figure 1 comparison is made between calculated and test results for a 1,500-kw 1,500 volt railway generator. Both the field and armature currents are shown. The calculations are made considering arcing at the brush edges such that the equivalent brush position is $1^1/_4$ brush widths from the quadrature axis. The machine is equipped with flash barriers whose location determines this allowance. Figure 2 is plotted to show the effect of shifting the armature magnetomotive force, and shows that if arcing is neglected the short-circuit peak current is 36 per cent higher, and the field current 26 per cent lower. It is obvious from Figure 3 that severe arcing will occur. There is little commutating pole flux at the moment of peak current. Therefore, the current in the brush face shifts drastically to the lee edge and causes the brush contact voltage to reach roughly 29 volts to accomplish commutation at rated load. Under heavy overload current, there is more than sufficient voltage for severe arcing.

In spite of these circumstances, the calculations are made for other machines taking the brushes to be centered on the quadrature axis with the center of current at the lee brush edge. This is the general condition similar to the distribution shown by Figure 3. Practical calculations can be made on this basis with good results, as Figures 4 and 5 show. Machines, in general, do not have the high voltage and high armature reactance peculiar to the machines of Figure 1.

Figure 4 is for another railway generator, 750-kw 1,500-volt, having brushes 26 per cent narrower than for the machine of Figure 1 and about 51 per cent less armature reactance voltage at rated load (without commutating pole flux). The calculated results show no large error in this case. The test oscillogram shows about 16 times normal current which is high for d-c machines in general and is caused chiefly by the 150-per-cent 2-hour load rating and 300-per-cent temporary load rating, for which the machine is designed. Hence unit current for the machine is an underrating relative to machines in general.

Both of these machines have pole face windings unlike the machine of Figure 5, which has interpole windings only. The latter machine is rated 300 kw, 250 volts, and shows slightly less than nine times rated current for the maximum. The machine has a series winding of one turn per pole which is taken into account in the calculation of field current. On this test the circuit breaker and cables have inductance of 0.26 per unit and resistance of 0.009 per unit (measured). These factors are included in the calculations.

In the calculations for Figures 1 and 4, the per-unit inductance of the machines is taken as 0.15 and for Figure 5 as 0.60 (see Table I). Winding resistances are measured values corrected to 75 degrees centigrade. Comparison of calculations and tests show good agreement of time constants.

These results, and calculations for other large motors and generators, show maximum currents from approximately 7 to 11 times rated current for machines with interpole windings only, and up to 14 for normally rated machines with both interpole and pole face windings. Such currents are most unlikely to occur in service, as faults without arcing or of zero resistance are rare. Maximum machine torque, in general, is less than 12 times rated torque (shown calculated and plotted for example in Figure 4). This is not difficult to take care of with safe margins in constructing machines. The weakest part is usually the coupling hub and key.

Discussion of Results

Fault Resistance

The formulas do not include fault resistance as such. When this exists it is included with the winding resistance of the machine. The effect of fault inductance can be allowed for fully by including it in the circuit inductance. Arcing faults are evaluated by allowing an average arc impedance or by calculating the results for "dead" short circuit and judiciously tempering the results.

Table I. Tests and Calculations to Determine C_x for Large Machines

Poles	Kw	Rpm	Volts	Test*	Calculated	Equation 1†	C_x
6	75	257	250	—	0.0062	0.0062	0.6
6	100	225	250	—	0.0042	0.0053	0.6
6	125	225	250	—	0.0043	0.0043	0.6
8	150	200	250	—	0.0020	0.0030	0.6
8	200	200	250	—	0.0023	0.0022	0.6
8	250	200	250	—	0.0018	0.0018	0.6
8	250	185	250	—	0.0017	0.0019	0.6
8	300	200	250	—	0.0016	0.0015	0.6
8	500	300	250	—	0.0008	0.00060	0.6
10	300	150	250	—	0.0016	0.0016	0.6
10	500	150	250	—	0.0011	0.00096	0.6
10	470	150	250	0.000958	—	0.00102	0.6
6	500	400	600	0.000700	—	0.00086	0.15‡
8	1,250	750	600	0.000132	—	0.00014	0.15‡
8	750	600	250	0.00020	0.00020	0.00020	0.6

* Test data are the a-c impedance at 25 cycles (checked at 60 cycles) measured between line terminals with the resistance of brush and brush contact short-circuited by copper jumpers and normal direct current in the shunt field winding.

† Equation 1: $L = \frac{E}{I}\frac{10.1C_x}{(\text{rpm})P}$

‡ Machines have pole face windings.

Self-Excitation and Compound Excitation

Self-excited machines on "dead" short circuit have negligible sustained field current and the same formulas applying, except that I_a and I_f, the sustained currents, are zero.

Compound wound machines give the same peak current, exciting flux, and torque as the basic separately excited machines. However, the peak field current is increased by an increment equal in per-unit terms to the ampere turns of the compounding winding per pole, divided by unit field ampere turns. The sustained field current is the same as in the separate or self-excited cases, respectively, but the sustained flux and armature current are different. Although the method can be extended to include these conditions, special formulas are not presented. Peak values are most important and are given by formulas 1, 2, and 3 for all types of machines except series wound. Occasionally, the method will need to be extended for compound wound machines when knowledge of sustained torques is necessary to study stresses in shafting, assuming that the field circuit instead of the armature circuit is interrupted and that overload relays are slow in acting or time must be allowed for dissipation of field energy to prevent excessive voltage. In any case, the reliability of the steady state results is affected by flashing between commutator segments creating multiple short-circuit current paths. In this connection some observations about flashing are important.

Flashing

A distinction is made between flashing current and arcing current. Flashing current is current flowing to or from the brush on the lee side having opposite direction of flow with respect to the arcing current flowing between brush and commutator on the same side. This is illustrated by Figure 6, making it clear that arcing at the lee brush edges cannot be carried around by the commutator to cause flashing current and actually is extinguished by reversal of the bar-to-bar voltage as the bars move away from the brushes. However, ionization of the air allows the armature electromotive force to strike new arcs across the commutator micas resulting in flashing current. Even with "dead" short circuit, there is substantial voltage around the commutator surface sufficient to cause flashing. Between brush arms this voltage is equal to the voltage across the commutating pole and pole-face windings plus the brush contact voltage. The latter voltage includes voltage equivalent to the a-c reactance voltage in the armature. The sum of these voltages is a substantial part of the voltage existing before short circuit.

These observations make evident the value of flash barriers[3] which sweep the ionized gas off the commutator and interfere with the flash path. They indicate that appreciable time is required to ionize the air and distribute it sufficiently to start flashing. Speaking of time in terms of 60 cycles per second, the current may rise to its peak in less than one cycle (see Figures 1, 4, 5). However, one-half cycle at machine frequency is taken for rotation of the armature through one pole pitch to carry ionized gases to the next brush arm. Most large d-c machines have frequencies less than 60 cycles so that breakers starting to interrupt current in one-half cycle (based on 60 cycles) should and actually do prevent flashover. Such breakers together with flash barriers give machines maximum protection.

BRUSH ARCING

Apart from flashing, arcing at the lee brush edges continues as the commutator bars leave the brush until current reversal in the respective coils is completed. The bar-to-bar voltage supports this arcing until polarity reverses some distance from the brush. This results in a shift of the armature magnetomotive force producing a component in the direct axis, increasing the field current, and shifting more field flux to the leakage paths reducing the armature electromotive force. If the amount of shift can be estimated, the effect may be included fully as provided for in Appendix I. However, in practice it is recommended that calculations be made assuming that the center of current flow between brushes and commutator is effectively at the lee edge of the brushes, knowing that arcing will reduce the current as, for example, in Figure 2. In general, the calculations will be pessimistic but not far wrong, as is shown by comparison of results with tests. Figure 3 illustrates the concentration of current at the lee edge of the brushes when there is negligible commutating pole flux. In Appendix IV it is shown that the effective brush contact resistance is the same with or without arcing so long as the bar-to-bar voltage, if appreciable, is considered consumed in the arcs. On this basis, brush arcing does not affect the circuit resistance materially, having its chief effect as a result of armature magnetomotive force shifted into the direct axis. As a basis for judgment in particular cases, it is helpful to understand the voltage required to support arcing.[3]

TRANSIENT AND STEADY STATE RESISTANCES

The formulas may be expressed in a form similar to the well-known formulas for synchronous machines[4,5] by letting $I_a' = 1/r_d'$ and $I_a = 1/r_d$, where r_d' and r_d are, respectively, transient and steady state effective resistances for short circuit with unit voltage on the machine. These resistances may be compared with the reactances x_d' and x_d, respectively, of synchronous machines; r_d' and r_d are affected by saturation and commutation and are variables. Unlike x_d', r_d' is not a definite coefficient widely useful in problems of transient voltage stability of generators or transient speed stability of motors. Generally, in d-c machines the armature magnetomotive force is held in quadrature with respect to the field magnetomotive force, and the armature and field windings are independent. Hence only winding resistance and self-inductance are effective in transient problems other than short-circuit performance. Of course, mutual inductance between series and shunt field windings always must be accounted for.

Appendix I. Derivation of Current and Torque Formulas

Consider for the moment a machine operating at normal voltage without load to be suddenly short-circuited, with no external resistance, and with constant speed maintained. The armature current rises very rapidly, but the rise of flux in the commutating poles is relatively slow because of the solid magnet frame. This flux rises fast in the skin of the frame, but, as saturation forces it to penetrate deeper into the solid steel, its rise is slowed. Accordingly, when maximum armature current is reached, the machine commutates substantially without benefit of commutating pole flux. Therefore, there is negligible compensation for the a-c reactance voltage in the armature winding, and an equivalent d-c voltage drop shows up at the brush contacts. This voltage is more than directly proportional to ordinary reactance voltage, as in a machine without commutating pole flux the current collection is crowded to one side of the brush faces with an effective reduction in width of brush face. The consequent change in wave shape of the current reversal in the armature conductors increases the reactance voltage. The equivalent direct voltage is determined and added to the *RI* drop of the windings and this sum equated to the electromotive force of the armature, giving an equation for the sum of the voltages in the armature circuit.

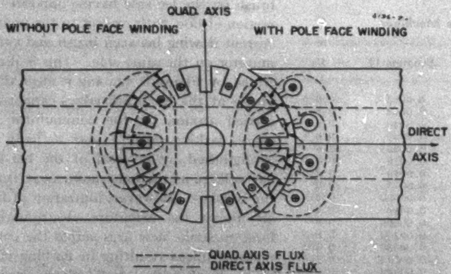

Figure 7 (above). Definition of axes and flux paths in machines with and without pole face windings

Figure 8 (right). Calculation of net direct axis flux resulting from combined direct and quadrature axes magnetomotive forces

See Table II for calculated data

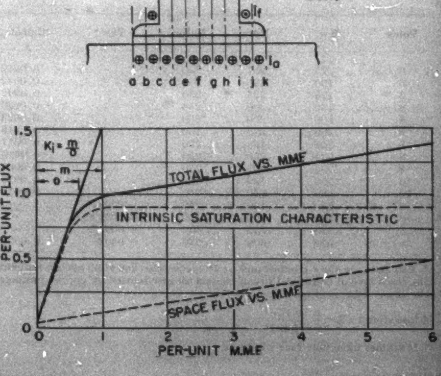

Table II. Calculated Data for Figure 8

		Net Mmf Acting on Section											Flux of Each Section											
I_a	I_f	a	b	c	d	e	f	g	h	i	j	k	a	b	c	d	e	f	g	h	i	j	k	ϕ_d
0	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
2.5	1	3.5	3.0	2.5	2.0	1.5	1.0	0.5	0	−0.5	−1.0	−1.5	1.21	1.17	1.12	1.08	1.04	1.0	0.75	0	−0.75	−1.0	−1.04	0.461
5.0	1	6	5	4	3	2	1	0	−1	−2	−3	−4	1.42	1.33	1.25	1.17	1.08	1.0	0	−1.0	−1.08	−1.17	−1.25	0.25
7.5	1	8.5	7.0	5.5	4.0	2.5	1.0	−0.5	−2.0	−3.5	−5.0	−6.5	1.63	1.50	1.38	1.25	1.12	1.0	−0.75	−1.08	−1.21	−1.33	−1.49	0.186
10	1	11	9	7	5	3	1	−1	−3	−5	−7	−9	1.83	1.67	1.5	1.33	1.17	1.0	−1.0	−1.17	−1.33	−1.5	−1.67	0.168
15	1	16	13	10	7	4	1	−2	−5	−8	−11	−14	2.25	2.0	1.75	1.5	1.25	1.0	−1.08	−1.33	−1.58	−1.83	−2.08	0.168

Although the build-up of the commutating pole flux is delayed, this is not so for the slot, end, and gap fluxes set up by the armature current in the laminated armature core and main pole shoes. These fluxes and what happens to the exciting flux are illustrated by reference to Figures 7 and 8. Figure 7 defines the direct and quadrature axes and shows the paths of flux under consideration. Figure 8 shows the pole of a machine without pole face windings and for simplicity with a uniform air gap. For illustration, the ampere turns of rated armature reaction at each pole tip are taken equal to the field ampere turns for the gap and teeth at normal voltage with no load. Then the magnetomotive force of armature reaction is divided into ten steps under the pole (not necessarily slots). At five times normal load amperes ($I_a=5.0$), the magnetomotive force per step is equal to the no-load normal-voltage gap and teeth ampere turns. Using per-unit terminology, this is unit magnetomotive force numerically equal to one. For any direct axis magnetomotive force applied to the gap and teeth, exciting flux (direct axis flux) and cross flux (quadrature axis flux) can be determined if the saturation curve for each section of the air gap is drawn as in Figure 8. Very large magnetomotive forces have to be handled, and this is done by fixing the slope of the upper part of the curve equal to that of an air gap whose length is equal to the normal air gap plus the depth of tooth. The basis for this is the well-established fact that above the density of "intrinsic" saturation limit (about 125,000 lines per square inch), the teeth act as though they were nonmagnetic to further increase in flux. In Table II calculated results for the direct axis flux (net exciting flux) are tabulated for direct axis magnetomotive force of 1, 2, 3, and 4 per unit and cross axis magnetomotive force of 0, 2.5, 5, 7.5, 10, and 15. These results are plotted in Figure 9 showing the relation between exciting flux field current and armature current.

To express the relationships by mathematical formula, the flux is divided into components. One of these is the flux existing in the gap and tooth space, taking the teeth to be nonmagnetic. This is the "space" component or "space" flux. The other component is the flux carried by the teeth at the intrinsic saturation limit. This is the "iron" component or "iron" flux. In calculating this with tapering teeth, the average width of the teeth is used. It is only necessary to calculate the flux for a single tooth and divide by the slot pitch to determine the "iron" flux per inch of pole arc. From the geometry of Figure 10, the following results are apparent:

$$\text{Quadrature axis (cross) iron flux} = \frac{b}{p_a}\phi_t$$

$$= \frac{c}{p_a}\phi_t \qquad (19)$$

where

ϕ_t/p_a = iron flux per inch of pole arc (= iron flux per tooth divided by slot pitch)

$$\text{Direct axis (exciting) iron flux} = \frac{a\phi_t}{p_a}$$

$$= \frac{M_f'p_p}{M_a'p_a}\phi_t \qquad (20)$$

where

p_a = pole arc (inches)
p_p = pole pitch (inches)
M_f' = peak transient field magnetomotive force
M_a' = simultaneous peak transient armature magnetomotive force

It is important to observe from Figure 10 that the portion of the arc (dimension *a*) which carries the net direct axis flux is acted on by magnetomotive force greater than that required for the intrinsic saturation limit. Sections of the pole arc *b* and *c*, where the iron may not be at this density, contribute only to the cross flux.*

The total direct axis (exciting) flux is determined by adding the space flux to the iron flux.

$$\text{Space flux} = M_f'\rho\frac{\phi_1 K_t}{M_{f1}} \qquad (21)$$

where

ϕ_1 = flux producing armature electromotive force equal to rated voltage at base speed, no load
M_{f1} = field ampere turns for rated voltage, no load
K_t = ratio gap and tooth ampere turns to gap ampere turns at flux ϕ_1
ρ = ratio of permeance of air gap with nonmagnetic teeth to that of the normal air gap

With pole face windings, the quadrature axis flux in the gap and teeth is as shown in Figure 7. The combined magnetomotive force of the armature and pole face windings is distributed across the pole arc as shown in Figure 11A. Its distribution across the arc between pole face bars is shown in Figure 11B where the magnetomotive force is that acting between the two windings from top of pole face copper to top of armature copper, rather than from bottom of pole face slot to bottom of armature slot as in Figure 11A.

The method used for determining the direct axis flux and the corresponding electromotive force under the influence of the strong quadrature flux is the same as for machines without pole face windings, except that the pitch between pole face slots is substituted for the pole arc and the magnetomotive force of the pole face winding is added to the armature magnetomotive force over the pole face slot pitch. Distortion of the flux distribution corresponding to the magnetomotive force distribution of Figure 11B exists at the top of the teeth because of criss-cross flux in the air gap and tooth-tip leakage flux. There is very little distortion at the bottom of the slots if the pole face magnetomotive force is well distributed and equal to the armature magnetomotive force. Magnetomotive force distributed as in Figure 11A is likely to be insufficient to saturate the teeth full depth, and in some cases the magnetomotive force acting on the tooth tops as in Figure 11B may be below that required to reach the intrinsic saturation limit of the iron over some part of dimension *a*. However, even in these special cases, negligible error in the results is caused by calculating the "iron" flux at the intrinsic saturation limit in the mean section of the armature teeth and dividing it into quadrature and direct axis components by the magnetomotive force ratio shown in Figure 11B. Then the direct axis iron flux is

$$\phi_{id} = \frac{a\phi_t}{p_b}$$

$$= \frac{M_f'p_a\phi_t}{p_b\left(\dfrac{p_aM_a'}{p_p}+M_b'\right)} \qquad (22)$$

This method takes care of machines with incomplete compensation and with varying number of pole face slots. If the pole face winding is removed, p_b becomes p_a and the formula reduces to equation 20.

The formula for space flux is the same as equation 21. Because the pole face slot tips saturate, ρ is the ratio of permeance of normal air gap to that with nonmagnetic armature teeth plus nonmagnetic pole face tooth tips.

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* It will be noted from Figure 10 that the average magnetomotive force acting on section *a* is p_aM_a'/p_p. When the flux caused by this magnetomotive force is derived for this section alone, the answer is identical with equation 20 plus equation 21. However, the factor ρ is derived to include with the "space" flux the fringing flux outside the pole arc. In this way, the fringing flux is not neglected.

From equations 21 and 22, the total exciting flux at the instant of maximum current is

$$\phi_d' = \frac{M_f' p_a \phi_t}{p_b\left(\dfrac{p_a M_a'}{p_p} + M_b'\right)} + M_f' \rho \frac{\phi_t K_t}{M_{f1}} \tag{23}$$

This has the unknowns ϕ_d', M_f', M_a', and M_b'. There are also the following equations holding simultaneously:

$$M_f' = \frac{M_{f1}}{I_{f1}} I_f' \text{ per unit} \tag{24}$$

where

I_f' = maximum field amperes
M_{f1} = base field magnetomotive force
I_{f1} = corresponding base field amperes

Similarly,

$$M_a' = \frac{M_{a1}}{I_{a1}} I_a' \text{ per unit} \tag{25}$$

$$M_b' = \frac{M_{b1}}{I_{a1}} I_a' \text{ per unit} \tag{26}$$

where

M_a' = maximum armature magnetomotive force
M_{a1} = armature magnetomotive force at rated amperes
I_{a1} = rated armature current
M_b' = maximum pole face winding magnetomotive force
M_{b1} = pole face winding magnetomotive force at rated armature current
I_a' = maximum armature current

Also, since the field linkages immediately after short circuit at the instant of peak current will be equal to those immediately before short circuit in accordance with the constant linkage theorem,

$$\phi_d' + n_f \frac{\phi_1}{I_{f1}} I_f' = \phi_o(1+n_f) \tag{27}$$

where ϕ_o is the initial field flux producing armature electromotive force. Since both sides of the equation are multiplied by the same number of turns to obtain linkages, this multiplier is omitted and the flux equation written directly. The term n_f is the fraction of ϕ_o which when multiplied by the same number of turns represents the remainder of the total field linkages, being that part which is caused by leakage flux not creating armature electromotive force.

The final necessary simultaneous equation is

$$I_a' = \frac{(\text{emf})'}{r_w + r_b} = \frac{1}{r_w + r_b} \phi_d' \frac{V}{n_1 \phi_1} \frac{60}{2\pi} \frac{d\theta}{dt} \tag{28}$$

where

r_w = resistance of windings (plus fault if necessary)
r_b = resistance of brush contact (equivalent to a-c reactance voltage of armature winding)
V = armature electromotive force (= rated voltage)
n_1 = rpm (= rated base speed)
ϕ_1 = flux producing unit electromotive force at unit speed
θ = angular displacement in radians
t = time in seconds

The paper is limited to the case of constant speed on the assumption that the drop in speed at the instant of peak current is negligible in all cases. For the sustained conditions, the results are limited to cases where the speed is maintained by the driving machine or reverse power flow to the driven machine.

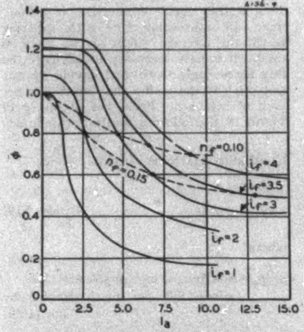

Figure 9. Relation between flux, armature current, and field current for machine represented by Figure 8

Circles and dashed lines connecting them are from equation 27

The equations give the following solutions:

$$I_a' = \frac{K_2\phi_o(1+n_f) - RK_1}{2R(K_2+n_f)} + \sqrt{\frac{K_1(1+n_f)\phi_o}{R(K_2+n_f)} + \left(\frac{RK_1 - K_2\phi_o(1+n_f)}{2R(K_2+n_f)}\right)^2} \tag{29}$$

$$I_f' = \frac{\phi_o(1+n_f)I_a'}{K_1 + (K_2+n_f)I_a'} \tag{30}$$

$$T' = \phi_d' I_a' = (K_1 + K_2 I_a')I_f' \tag{31}$$

$$P' = (I_a')^2 R \tag{32}$$

where, in terms of per-unit quantities,

$$K_1 = \frac{\phi_1 p_a p_p}{p_b(p_a M_{a1} + p_p M_{b1})} \tag{33}$$

$$K_2 = \rho K_1 \tag{34}$$

$$R = r_w + r_b \tag{35}$$

The steady state results are obtained by solving the equations, excluding equation 27, which does not hold for steady state conditions. I_f, the sustained field current, is substituted for I_f'. For separately excited machines, I_f is the same as the initial field current I_{f0}. For self-excited machines on "dead" short circuit, $I_f = 0$. The solution for the steady state gives

$$I_a = \frac{K_2 I_f}{2R} + \sqrt{\frac{K_1}{R} I_f + \left(\frac{K_2 I_f}{2R}\right)^2} \tag{36}$$

$$T = (K_1 + K_2 I_a) I_f \tag{37}$$

The effect of shift of the axis in armature magnetomotive force mentioned under "Brush Arcing" can be included by replacing M_f' in equation 23 with M_d', where

$$M_d' = \frac{M_{f1}}{I_{f1}} I_f' - \frac{2bP}{B} \frac{M_{a1}}{I_{a1}} I_a' \tag{38}$$

and using this equation in place of equation 24

b = brush width in number of segment pitches
P = the number of poles
B = total number of commutator segments

This gives the formulas presented in "Formulas for Current and Torque." Any shift from the quadrature axis measured in equivalent number of commutator segments can be accounted for by replacing b by twice the number of segments of shift.

To develop complete current-time formulas, time constants for the armature and field circuits are developed in Appendix II. The reciprocals of these time constants give the corresponding decrement factors. Then, if

σ_a = decrement for armature circuit
σ_f = decrement for field circuit
I_a' = peak armature current
I_a = sustained armature current

The armature current has the following components:
Rise:

$$I_a'(1 - \epsilon^{-\sigma a t})$$

Decay:

$$-(I_a' - I_a)(1 - \epsilon^{-\sigma f t})$$

The time constants are so dissimilar that the components may be put together directly with negligible error. Therefore,

$$I_a = I_a - I_a' \epsilon^{-\sigma a t} + (I_a' - I_a)\epsilon^{-\sigma f t} \tag{39}$$

Similarly,

$$i_f = I_f - (I_f - I_f')(\epsilon^{-\sigma f t} - \epsilon^{-\sigma a t}) \tag{40}$$

Appendix II. Time Constants

The field time constant is

$$T_f = \frac{L_f}{R_f} = \frac{N\phi_1 K_t P(1+n_f)10^{-8}}{V_f} \text{ seconds} \tag{41}$$

where N is the number of turns per coil, and V_f is the field voltage including rheostat. ϕ_1 is in maxwells.

The armature time constant is

$$T_a = \frac{L_a}{r_w + r_b}$$

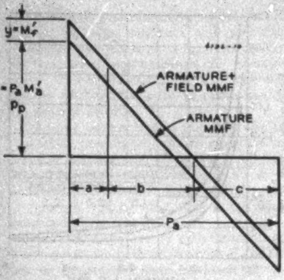

Figure 10. Distribution along air gap showing net exciting magnetomotive force over section a

$$c:(x-y)=(a+b):(x+y)$$
$$b=c,\ P_a=a+b+c$$
$$\text{Therefore, } \frac{a}{P_a}=\frac{y}{x}=\frac{M_f'P_p}{M_a'P_a}$$

L_a = armature circuit inductance $=\dfrac{k_1T^2p_q}{a^2P}$, neglecting slot and end leakage (42)

where

T = total number of armature turns
p_q = total gap permeance for armature magnetomotive force
a = number of circuits in armature winding
P = number of poles

Also,

$$\frac{E}{I}=\frac{k_2T\,(\text{rpm})\,p_d\,(\text{field mmf})\,P}{aI} \tag{43}$$

where

E = rated volts
I = rated amperes
p_d = excitation permeance
rpm = rated base speed

Let field magnetomotive force be proportional to armature reaction:

$$\text{Field magnetomotive force}=\frac{k_3TI}{aP} \tag{44}$$

and let p_d be proportional to p_q. Then, substituting equation 44 in equation 43 and substituting the result in equation 42,

$$L_a=\frac{E}{I}\,\frac{C_x}{P(\text{rpm})} \tag{45}$$

This indicates that under the assumptions L_a is equal to unit resistance divided by machine frequency and multiplied by a coefficient C_x. The value for C_x, representative of some d-c machines, has been determined from test data and calculations shown in Table I. Using this quantity it is possible to estimate the inductance and, therefore, to specify the approximate time constant. When resistance is expressed in per unit,

$$T_a=\frac{19.1C_x}{(r_a+r_b)(\text{rpm})P}\ \text{seconds} \tag{46}$$

Saturated Time Constants

The rise of current is governed at first by the normal armature time constant, but finally any change in flux linkages of either the armature or field circuits for a given change in their respective currents is substantially in proportion to the space permeance illustrated in Figure 8. Calculations similar to those of Figure 8 demonstrate this. Therefore, the saturated time constants are

$$T_{as}=\frac{\rho+n_f}{1+n_f} \tag{47}$$

and

$$T_{fs}=\frac{\rho+n_f}{1+n_f} \tag{48}$$

Equation 47 is an approximation for the armature winding taking the portion of the flux linkages not affected by saturation to be the same percentage of total linkages as is the case for the field winding.

Appendix III. Permeance Ratio ρ

The normal air gap permeance is (omitting numerical coefficients)

$$P_n=\frac{A}{gK_\lambda C_sC_d} \tag{49}$$

where

A = area of gap = pole arc × pole length
g = average length of air gap
K_λ = ratio flux under pole arc to total gap flux, Figure 12 (fringing factor for boundaries of pole face)
C_s = slot fringing factor, Figure 14
C_d = duct fringing factor, Figure 13

The space permeance with nonmagnetic teeth is

$$p_n'=\frac{A}{(g+d_s)K_\lambda'} \tag{50}$$

where

d_s = depth of slot
K_λ' = 0.85 (very nearly constant for very large gaps)

Therefore,

$$\rho=\frac{p_n'}{p_n}\,\frac{gK_\lambda C_sC_d}{0.85(g+d_s)} \tag{51}$$

In cases with pole face windings, add to d_s the depth from pole face to top of copper in pole face slots.

Appendix IV. A-C Reactance Voltage of Armature and Equivalent Brush Contact Resistance r_b

Assuming that the machine operates momentarily without commutating pole flux because of the extremely rapid rise of armature current, all of the reactance voltage appears as an equivalent brush contact voltage. However, the normal reactance voltage is increased by crowding the current to the edge of the brush, in turn increasing the rate of current reversal in the coil and the reactance voltage. Under these circumstances the equivalent brush drop voltage is arrived at in the following way.

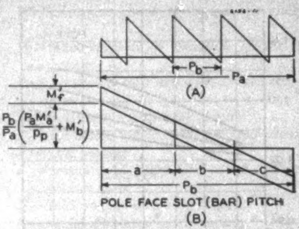

Figure 11. Difference in combined magnetomotive force distribution acting at top and bottom of slots in machines with pole face windings

A—Acting across space from bottom of pole face slot to bottom of armature slots ($=M_f'+M_a'-M_b'$)

B—Acting across space from top of pole face copper to top of armature copper

If a reactor of negligible resistance is suddenly connected in parallel with a resistor which is part of a circuit containing a driving voltage and a steady circulating current, the current will shift from the resistance path to the inductive path. While this shift is happening, the energy lost in the resistor is equal to the change in energy stored in the reactor. The resistor is analogous to the net brush contact resistance and the reactor to the armature coil. The energy change in a coil can be calculated and, since the number of coils rotating past the brush per second is known, the equivalent watts at the brush contact can be determined. It will be assumed that the *average* reactance voltage per coil E_R is known or is measured as in Appendix V. Then the inductance per coil is

$$L_c=\frac{E_Rb}{I_bf_s}\times10^8 \tag{52}$$

where

I_b = amperes per brush arm
f_s = segments per second (= number of segments times revolutions per second)
b = brush overlap (= width brush face in number of segment pitches)

The energy change per coil is

$$\Delta\epsilon={}^1/_2L_cI_b^2$$
$$=\frac{E_RbI_b}{2f_s} \tag{53}$$

The equivalent watts are

$$f_s\Delta\epsilon=\frac{E_RbI_b}{2} \tag{54}$$

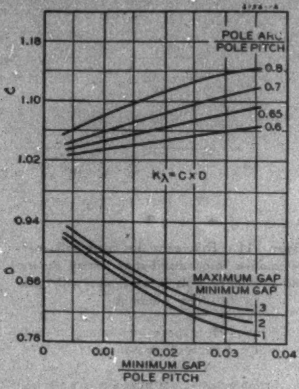

Figure 12. Fringing factor for boundaries of pole face K_λ

and the equivalent brush contact resistance for plus and minus brushes is

$$r_b = \frac{E_R b}{I_b} \tag{55}$$

This is the minimum brush resistance, assuming all the brush contact voltage counteracts reactance voltage with constant contact resistance. If L_e is nearly constant, this is also the minimum contact resistance for either arcing contact or for brush contact resistance varying with current. But in all cases, all the voltage produced is needed to counteract reactance voltage. The true voltage is more than this minimum because it is known that arc voltage consists of a "constant" base voltage plus a variable voltage depending on arclength and current.[8] The "constant" base voltage is surplus and cannot counteract reactance voltage as only the voltage difference between segments can do this. Moreover, the author has found the resistance of carbon brush contacts at high density to be nearly equivalent to a base voltage plus a constant times the brush current. Undoubtedly, there is some surplus voltage in the brush contact with or without arcing, and, therefore, the contact loss of brushes under rated conditions assuming perfect commutation is added to allow for it approximately. The final result for r_b on a per-unit basis is

$$r_b = \frac{E_R b + V_c}{V} \tag{56}$$

where

V_c = approximate contact drop of the plus and minus brushes at rated load current density

V = rated voltage

Equation 56 is approximately true whether or not arcing is involved in commutation, so long as the effective self-inductance of the armature coils is not affected. A correction factor sometimes needed for equation 56 and an expression for the distribution of current in the brush face are derived by solving the basic equation of a coil undergoing commutation. This equation says that the reactance voltage is equal to the difference in brush contact voltage between the brush and the two segments to which the coil is connected or, from Figure 15,

$$L\frac{d}{dt}i_c = \frac{(Rt_b)}{(t_s)}t_s\frac{d}{dt}i_s \tag{57}$$

$$i_s = t_s\frac{d}{dt}i_c \tag{58}$$

from which

$$\frac{d}{dt}i_s - \frac{L}{Rt_bt_s}i_s = 0 \tag{59}$$

where

$RI_b = V_c$ = normal brush contact voltage

$\frac{Rt_b}{t_s}$ = brush contact resistance/segment

$L = E_Rt_b/I_b$

The solution to equation 59 is

$$i_s = A\epsilon^{pt}$$

where

$$\begin{aligned} p &= L/Rt_bt_s \\ &= E_R/RI_bt_s \\ &= \frac{E_Rf_s}{V_c} \end{aligned}$$

and A is found from

$$\begin{aligned} \text{Average } i_s &= \frac{1}{t_b}\int_0^{t_b} i_s\,dt \\ &= \frac{A}{t_b}\,\frac{(\epsilon^{pt_b}-1)}{p} \\ &= \frac{I_bt_s}{t_b} \end{aligned}$$

from which

$$i_s = \frac{E_R\epsilon^{pt}}{R(\epsilon^{pt_b}-1)} \tag{60}$$

This equation is plotted for a 1,500-kw 1,500-volt generator in Figure 3.

The watts loss in the brush contact is

$$\begin{aligned} \frac{1}{t_s}\int_0^{t_b} i_s^2\frac{(Rt_b)}{(t_s)}\,dt &= \frac{t_bE_RI_b}{2t_s}\times \\ &\quad \frac{(\epsilon^{2pt_b}-1)}{(\epsilon^{pt_b}-1)^2} \\ &= I_b^2R_b \end{aligned} \tag{61}$$

where

R_b = effective brush contact resistance

The total brush contact resistance is in per-unit

$$\begin{aligned} r_b &= 2R_bI_b/V \\ &= \frac{E_Rb}{V}\left(\frac{\epsilon^{2x}-1}{\epsilon^{2x}-2\epsilon^x+1}\right) \end{aligned} \tag{62}$$

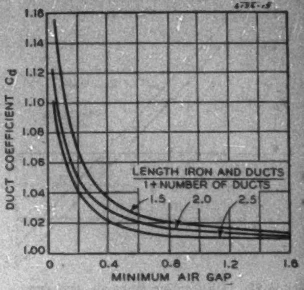

Figure 13. Duct fringing factor C_d

where

$$x = \frac{E_Rb}{V_c}$$

If x is small in some cases so that the bracketed factor is appreciably greater than one, the first term of equation 56 should be increased accordingly.

Appendix V. Constants From Test Measurements

In many cases the results may be used without reference to manufacturer's data.

ϕ_t may be taken as equal to 1.0 in large machines because the tooth density at normal voltage will approach 125,000 lines per square inch. For small machines ϕ_t may be as much as 1.25. An error does not affect the results seriously for peak armature current and torque as, in any case, the exciting flux is stoutly maintained unless the field leakage flux is unusually large ($n_f > 0.15$).

The term ρ may be determined by measuring the air gap and the depth of slot with gauge and rule. Then by using equation 51, taking $K_\lambda = 0.93$, $C_s = 1.1$, and $C_d = 1.04$, the value of ρ may be computed. These are representative values for large machines.

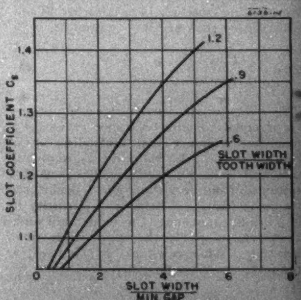

Figure 14. Slot fringing factor C_s

Average or representative values for K_1 and K_t for large machines with pole face windings are 4.0 and 1.3, respectively. Without pole face windings, K_1 averages about 1.0 with no change in K_t. The field leakage coefficient n_f may be taken as 0.12 (an average figure).

All these factors chiefly influence the induced field current required to maintain the field flux linkages, and when n_f is small they have relatively little influence on the peak armature electromotive force, current, and also torque. If n_f were zero, they would have no influence at all.

The remaining factor of major importance is the resistance r_b dependent on E_R. This can be determined from equation 13 and the relation

$$E_R = 6.38 \times AT \times 1/g \times l \times S \times 10^{-8} \qquad (63)$$

where

AT = net difference in the ampere turns at rated load of the armature and compensating windings determined by counting their respective turns from inspection
g = measured length of the commutating pole air gap and nonmagnetic shims
l = measured length of the armature core in inches
S = revolutions per second times the measured periphery of the armature in inches

Another method for estimating E_R is from the relation

$$E_R = \frac{C_r \text{ times } K_w}{d_a} \qquad (64)$$

where

C_r = reactance voltage coefficient averaging about 0.12 for large machines
K_w = rated kilowatts
d_a = armature diameter in inches

If a fluxmeter or ballistic galvanometer is available it may be connected to an exploring coil around the tip of the commutating pole to measure the flux, which when multiplied by f_s/b and by 10^{-8} will give E_R (roughly) for a half coil turn. Similarly, an exploring coil around the base of the main pole and another around the pole face can be used to measure n_f approximately.

Appendix VI. Nomenclature

AT = net difference between armature ampere turns and ampere turns of commutating pole plus pole face windings at maximum rated continuous current
b = width of brush contact face in commutator segments
B = total number of commutator segments
C_d = flux fringing coefficient for armature ducts (Figure 13)
C_r = reactance voltage coefficient (=0.12 for large machines)
C_s = flux fringing coefficient for armature slots (Figure 14)
C_x = average per-unit inductance of d-c machines (see Table I)
d_a = armature diameter in inches
d_s = depth of armature slot in inches
E_R = rated load average reactance voltage at speed for which short circuit is calculated
f_s = commutator segments per second passing brush arm
g = main pole average air gap in inches
I_a = steady state short-circuit armature amperes
I_{a1} = maximum rated continuous armature amperes
I_a' = maximum or peak transient short-circuit armature amperes
I_b = rated amperes per brush arm
I_{f1} = field amperes for rated voltage at base speed, no load
I_f' = maximum or peak transient field amperes on short circuit
K_1 = "iron" flux constant, equation 11
K_2 = "space" flux constant, equation 12
K_t = air gap plus tooth ampere turns divided by air gap ampere turns at rated voltage, base rated speed, no load
K_λ = pole arc air gap flux divided by total gap flux (Figure 12)
l = length of iron core portion of armature including ducts in inches
M_{a1} = total armature ampere turns per pole at maximum rated continuous current (equals current times total turns divided by circuits, divided by poles)
M_a' = peak transient armature ampere turns per pole on short circuit
M_{b1} = pole face winding ampere turns per pole at rated current
M_b' = maximum transient ampere turns of pole face winding per pole on short circuit
M_{f1} = field ampere turns at rated voltage, rated base speed, no load
M_f' = maximum transient field ampere-turns on short circuit
n = speed for which short circuit is calculated
n_1 = rated base speed in revolutions per minute
n_f = field leakage flux linkages divided by field linkages from flux generating armature electromotive force
N = number of field winding turns per pole
p_a = pole arc in inches
p_b = pole face slot pitch (average) in inches
p_p = pole pitch in inches
P = number of poles
r_b = equivalent resistance of brush contacts
r_w = resistance of machine windings
$R = r_w + r_b$ = total effective resistance of circuit
S = revolutions per second times armature circumference in inches
t = time in seconds
T = torque as function of time
T' = maximum transient torque on short circuit
T_a = time constant of armature circuit in seconds
T_{as} = saturated time constant of armature circuit
T_f = time constant of field circuit
T_{fs} = saturated time constant of field circuit
V = rated voltage
V_c = normal load brush and brush contact voltage drop
V_f = field voltage including rheostat
θ = angular displacement in radians
ρ = ratio permeances: air gap plus nonmagnetic teeth to normal air gap
σ_a = armature decrement factor (reciprocal of T_a or T_{as})
σ_f = field decrement factor (reciprocal of T_f or T_{fs})
ϕ_0 = flux before short-circuit generating armature electromotive force
ϕ_1 = flux per pole generating rated voltage at rated base speed
ϕ_d' = flux generating electromotive force immediately after short circuit

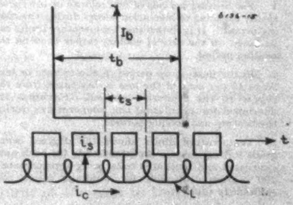

Figure 15. Circuit diagram for equation governing current reversal in armature coils and flow of current between segments and brush

ϕ_t = "iron" flux (=pole arc times mean section of tooth times 125,000 lines per square inch divided by slot pitch)

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Temperature Limits for Short-Time Overloads for Oil-Insulated Neutral Grounding Reactors and Transformers—II

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Synopsis: In part I, published in 1938,[1] the principal question considered was the effect of ventilating coils on temperature limits for ground faults greater than one minute, for which it was assumed that 160 degrees centigrade based on all heat stored in copper was the correct limit.

Since 1938 considerable new data have been obtained. This paper presents the new data and shows that:

1. For heating periods of one minute or less, practically all aging of insulation occurs during the cooling period. It is shown that temperature limits can be based on the rate of cooling rather than on the heating period.
2. In the heating-up period of one minute or less, 25 to 45 per cent of the loss is dissipated from the copper to the insulation and oil. Formulas are developed for calculating the temperatures during the heating and cooling periods.
3. Temperature limits above 160 degrees centigrade for one minute are permissible for both nonventilated and ventilated coils, being approximately 200 degrees centigrade for windings which cool sufficiently rapidly.

THE earlier paper[1] gave data on the heating of coil stacks with and without ducts, and served as the basis of selecting 160 degrees centigrade as a suitable limiting temperature for short-time rating of grounding devices using class A insulation. The standard for neutral grounding devices, AIEE Standard 32, is undergoing revision, and this paper is being presented to give data to serve as the basis of selecting suitable temperature limits.

The heating time intervals generally accepted for neutral grounding devices are ten seconds and one minute. For such short heating periods, the cooling time is more effective on the aging of the insulation than the heating time. Accordingly, the temperature limit for a 10-second coil can be only a few degrees higher than the limit for a 1-minute coil.

Previous calculations of the short-time heating of neutral grounding devices have been made on the basis of all heat stored in the copper. This practice results in calculated rises appreciably higher than actual, and, in reality, furnishes no experience at all in reference to the suitability of 160 degrees centigrade as a temperature limit.

In the transformer standard American Standards Association C-57, the temperature limit is 250 degrees centigrade with the calculations based on all the heat being stored in the copper, the same as for neutral grounding devices, and with full sustained voltage applied to the transformer. Again this gives no clue as to what temperature might be considered safe for neutral grounding devices because it is definitely known that the 250 degrees centigrade limit never is reached, nor even approached very closely.

Accordingly, the question of the selection of suitable limits for neutral grounding devices must be approached directly on its own merits. Several papers on aging have been published, and it is more or less generally agreed that the rate of aging doubles for a definite temperature increment. In a recent paper by F. M. Clark,[2] he shows that the increment of temperature for which doubling occurs is relatively small at 95 degrees centigrade and is practically 8 degrees centigrade for temperatures upward from approximately 115 degrees centigrade. This confirms the value originally reported by V. M. Montsinger.[3] Some authors have given values up to 10 degrees centigrade. However, the value of 8 degrees centigrade as the increment generally is accepted for transformer insulation.

In Clark's paper there are data indicating that 16 doubles in aging effect occur over the temperature range from 100 to 200 degrees centigrade, and it seems reasonable to use this rate of aging

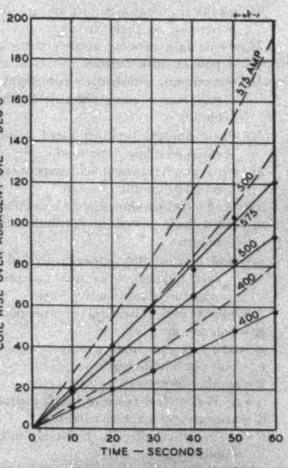

Figure 1. Curves of tested and calculated temperature rises during 60-second duration in 24-degree-centigrade oil

------- Calculated, all heat stored in copper only
——— Line through test points
Size of conductor, 0.435 by 0.095 inch

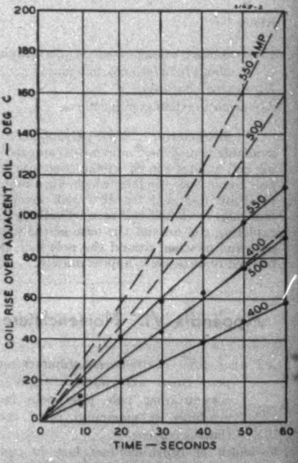

Figure 2. Curves of tested and calculated temperature rises during 60-second duration in 71-degree-centigrade oil

------- Calculated, all heat stored in copper only
——— Line through test points
Size of conductor, 0.435 by 0.095 inch

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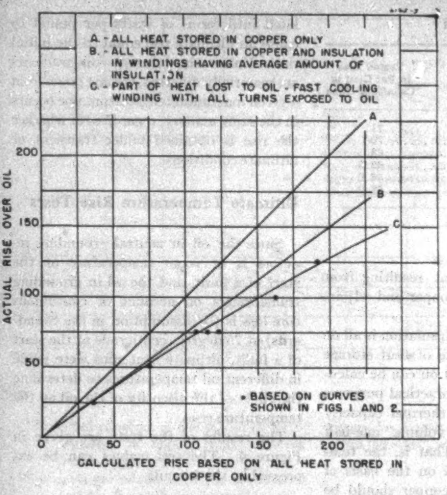

Figure 3. Curves of calculated temperature rise and hot-spot temperature rise under 1-minute fault condition

increase in estimating suitable temperature limits for neutral grounding devices. It should be borne in mind, however, that aging tests are tests of a mechanical nature in that loss of mechanical strength is measured. Such tests give no clue as to the actual length of life to be expected in transformers or similar devices. The aging tests, however, do furnish a means of making comparisons so that we can say that greater or less life can be expected at lower or higher temperatures. Further, the deviations in the results of life tests are quite large so that we must talk of relative aging in rather broad terms. The translation of the results of aging data into life estimates must be made cautiously and with common sense and judgment.

Quite obviously, the selection of the temperature limit will depend upon the answer to the question, "How much aging can be allowed for each operation?" This could be answered if we knew how many faults would occur per year. However, lacking that knowledge, it is necessary to make the best estimate possible on the basis of the data we have. After considerable investigation it appears that the temperature limit should be selected on the basis that the relative aging of a neutral grounding device during one year be the same as that of a power or distribution transformer during the same period, operating at rated load in the reference ambient temperature, and also that ten events (faults at full rated current) occur per year.

Since most of the aging occurs during the cooling of the coil, it appears quite obvious that the higher the rate of cooling, the higher can be the temperature limit. The cooling time varies directly as the thermal capacity associated with one pound of copper, varies inversely in an exponential manner as the watts per pound required to produce a given ultimate temperature rise, and is a direct function of the heating characteristics of the winding, that is, the temperature rise for one watt per sound. Obviously a multitude of limits might be established depending upon the correlation of these factors, but this does not seem desirable.

Designation of Windings Used in Neutral Grounding Devices

Since as much as 45 per cent of the heat escapes to the insulation and oil during a 1-minute fault in windings having cooling ducts between all coils, it is desirable to have some easy means of distinguishing between these windings and windings with no cooling ducts from which no heat escapes to the oil.

For purposes of identification in this paper the usual well-ventilated coils will be designated as "fast cooling windings" and the "no duct" or poorly-ventilated coils will be designated as "slow cooling windings," without consideration of grammatical correctness.

Heating Period Under Fault Condition

Fast Cooling Windings

In the past it has been assumed that for a short-time fault condition all heat is stored in the copper, that is, the thermal capacity of the conductor insulation was neglected. However, recent tests have shown that for windings having each conductor exposed on both edges to the oil, between 25 and 45 per cent of the copper loss is dissipated at the end of 60 seconds.

The tests referred to previously were made on horizontal disk coils having 0.435-by-0.095-inch conductors with a two-sides thickness of turn insulation of 0.019 inch. The radial build of the coils was 2.625 inches. The hot-spot temperature was measured by a thermocouple embedded under a spacer and between turns of the next to top coil. All temperature rises were based over the adjacent oil, 1.5 inches away from the coil containing the thermocouple.

The temperatures were read on an automatically balanced indicating type of potentiometer. The lag in temperature was quite small because the temperature started falling within one to two seconds after the current was removed. The direction of the current was reversed in alternate coils to reduce stray loss to a minimum. The measured loss was practically the same as the I^2R.

Figures 1 and 2 show curves of meas-

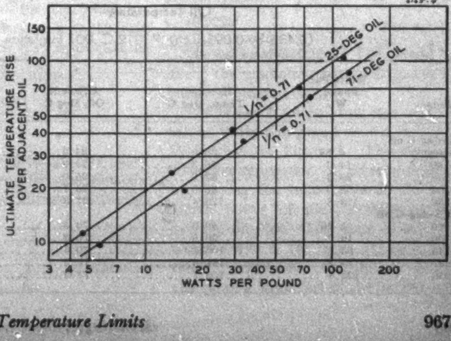

Figure 4. Ultimate hot-spot temperature rise versus watts per pound of disk coils

Disk coils have 0.435-by-0.095-inch conductors, 2.625 inches radial build, 3/8-inch oil ducts between coils, and are immersed in 25- and 71-degree-centigrade oil

Table I. Tested Rises in Per Cent of the Calculated Rises

Approximate Oil Temp, Deg C	Amp	Temp Rise at End of 60 Sec Over Adjacent Oil, Deg C: Calculated*	Temp Rise at End of 60 Sec Over Adjacent Oil, Deg C: Tested	Tested Rise in Per Cent of Calculated Rise
25	400	80	57	71
25	500	135	94	70
25	575	191	120	63
71	400	93	58	62.5
71	500	150	90	56.0
71	550	201	114	57.0

*On basis all heat stored in copper.

ured temperature rises, holding three different values of amperes for 60 seconds on coils immersed in approximately 24- and 71-degree-centigrade number 10-*C* oil. The dashed lines show the temperature rises that would have been attained if all the heat had been stored in the copper. The formula for calculating these rises is shown in Appendix I.

These tests show that the temperature rise of fast cooling coils under a 1-minute fault can be assumed to be approximately 65 per cent of the rise based on all heat stored in copper only. For the higher oil temperatures, the value is less than 65 per cent. Fifty-five to 65 per cent would be more nearly correct, so 65 per cent is conservative.

Figure 3 shows the relationship between calculated and actual rises in windings ranging from fast cooling to the slowest cooling.

Slow Cooling Windings

Under slow cooling windings may come

1. Windings having no oil ducts between individual coils.
2. Windings or coils heavily taped.
3. Windings having several layers between oil ducts.

For windings coming within the above classifications, it can be assumed that the temperature rise within a period of 60 seconds is close to that resulting from all heat stored in the copper and adjoining insulation.

Where the adjoining insulation is all on the conductors, the rate of heat storage in copper and its insulation can be calculated close enough for practical purposes by assuming that the thermal capacity of the insulation is "by volume" one-half that of the copper.[4] That is, the temperature rise calculated on the basis of all heat stored in the copper should be multiplied by the factor

$$\left(\frac{2a}{A+a}\right)$$

where

A = cross-sectional area of copper plus insulation
a = cross-sectional area of copper

Cooling Period Following a Fault

Because most of the aging of the insulation even in fast cooling windings occurs during the cooling period following a 1-minute fault as shown later, it is important to be able to calculate the cooling of various kinds of coils.

To calculate the cooling of a winding that has not reached ultimate conditions requires that the initial temperature rise (at start of the cooling period) be translated into terms of watts per pound to cause an ultimate rise equal to the initial rise. Unless the internal coil gradients are materially different, cooling down from a given maximum temperature rise occurs in the same manner, regardless of whether the rise is obtained under transient or ultimate conditions.

Ultimate Temperature Rise Tests

Since the oil in neutral grounding reactors is at room temperature at the start of a fault, and the oil in grounding transformers on account of continuous core loss is (by assumption in the Standards) at 75 degrees centigrade at the start of a fault, ultimate heat runs were made in different oil temperatures to determine the effect of the viscosity of the oil on the temperature rises.

The results of the tests are given in Figure 4. The test values can be expressed by the formula

$$\Theta = kW_c^{\frac{1}{n}}\mu^{0.25} \qquad (1)$$

where

Θ = temperature rise, degrees centigrade
k = constant
W_c = watts per pound of copper at ultimate temperature
μ = viscosity of the average temperature of the oil film expressed in centipoises
n = empirical exponent of temperature rise Θ versus W_c

Figure 4 shows the relationship between ultimate tested temperature rise and watts per pound of copper. For these curves, when viscosity is not factored $1/n$ = approximately 0.70. When viscosity is factored, $1/n = 0.75$ in the formula. Similar tests were made on other sizes of conductors and conductor insulation of two-sides thickness up to and including 0.073 inch. The tests made on one size of conductor and insulation are given in Table II. These values are plotted in Figure 4.

Cooling After Fault Condition

Where temperature rise varies directly with the loss, the cooling can be calculated by the formula

$$\Theta = \Theta_t\epsilon^{-t/B} \qquad (2)$$

where

Θ = temperature rise at any time t
Θ_t = initial temperature rise over oil
t = time of cooling in minutes
B = time constant, for time in minutes

$$B = \frac{\Theta_t C}{W_0}$$

Table II. Summary of Tests Made on Coils for Ultimate Temperature Rises in Different Oil Temperature

(0.435-by-0.095-inch) P .012 C .007 Insulation

Amp	Watts Per Pound, W_c	Hot-Spot Temp, Deg C	Adjacent Oil, Deg C	Hot-Spot Temp Rise Over Oil, Deg C Test
25-deg-C oil				
60	4.88	36.9	24.9	12.0
100	14.2	50.3	24.9	25.4
140	29.6	67.6	25.0	42.2
202	68.0	98.5	25.7	72.8
250	113.8	129.4	27.3	102.1
70.5-deg-C oil				
60	5.66	80.0	70.5	9.5
100	16.3	90.8	70.8	19.5
140	33.6	106.3	70.5	35.7
207	79.8	137.5	70.2	67.1
250	122.4	156.7	71.2	85.5

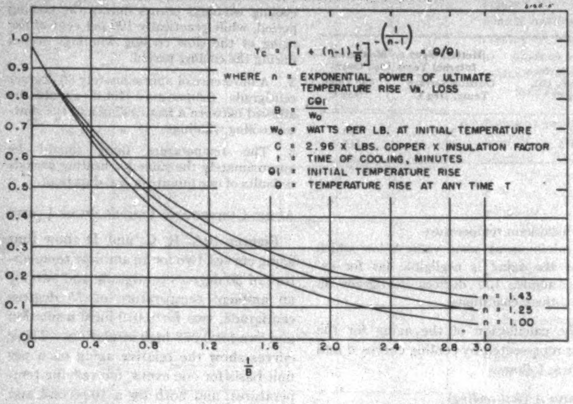

Figure 5. Cooling curves for windings in which loss = $k\Theta^n$, where n varies from 1 to 1.43

C = thermal capacity of one pound in watt-minutes of copper plus adjacent insulations

W_0 = watts per pound of copper required to produce an ultimate rise equal to Θ_t

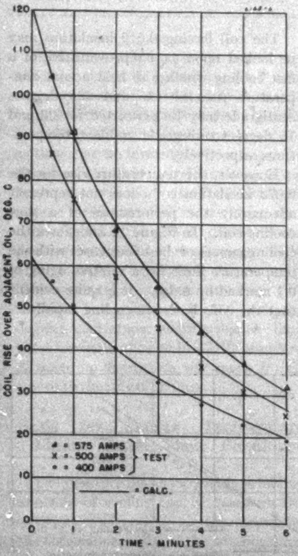

Figure 6. Curves of tested and calculated cooling after 1-minute heating tests

Tests made in 24-degree-centigrade oil
Size of conductor, 0.435 by 0.095 inch
Solid line calculated by equation 3

However, when temperature rise varies with loss raised to a power in the order of 0.8 or less, equation 2 does not apply, and an equation of the following form applies:

$$Y_C{}^* = \left[1+(n-1)\frac{t}{B}\right]^{\frac{-1}{n-1}} \qquad (3)$$

where

$Y_C = \Theta/\Theta_t$

n = exponent of ultimate temperature rise Θ versus loss

$$B = \frac{C\Theta_t}{W_0}$$

Figure 5 gives a plot of Y_C versus t/B when $n = 1.43$, that is, 1/0.7, 1.25, and 1.0.

Example of Calculated Cooling

Assume a temperature rise of 67 degrees centigrade in 71-degree-centigrade oil. At the end of an overload where 67 degrees centigrade is the initial rise for the cooling cycle, the copper loss by Figure 4 is 80 watts per pound.

The value of C (thermal capacity of copper and insulation) for (0.435 inch by 0.095 inch) conductors with 0.019-inch two-sides thickness of turn insulation, is by previous statement

$$C = 2.96\left(\frac{0.0518+0.0408}{2\times0.0408}\right) = 3.36$$

Other tests not reported in this paper show that the spacers add approximately 10 per cent to the thermal capacity. That is,

$$C = 3.36\times1.1 = 3.7$$

Therefore

$$\frac{1}{B} = \frac{80}{3.7\times67} = 0.322$$

A comparison between the tested and calculated values (by equation 3) is shown in Table III.

Table III

t	$\frac{t}{B}$	Y_C	Temp Rise, Deg C $Y_C\Theta_t$ Calculated	Temp Rise, Deg C Test
0	0	1.0	67.0	67.0
1/2	0.161	0.86	57.5	57.0
1	0.322	0.735	49.2	49.0
2	0.644	0.565	38.0	38.5
3	0.966	0.445	30.0	30.8
4	1.288	0.353	23.6	25.4

Figures 6 and 7 give the tested and calculated cooling curves for six different tests made in 25- and 71-degree-centigrade oil.

Cooling of Windings Having Few or No Cooling Ducts

Tests made on a typical poorly ventilated coil stack consisting of a number of coils of 3.625 inches radial build, assembled without cooling ducts and having several thermocouples embedded between the coils, gave maximum ultimate internal temperature rises as shown in Table IV. The wire size was 0.400 by 0.040 inch with 0.021-inch two-sides thickness of turn insulation.

Then

$$O = 2.96\left(\frac{A+a}{2a}\right)$$
$$= 2.96\left(\frac{0.0416}{0.032}\right)$$
$$= 3.85$$

and

$$\frac{1}{B} = \frac{9.8}{3.85\times115}$$
$$= 0.0222$$

The value of 0.0222 for $1/B$ should be about the minimum that would occur in poorly ventilated coils.

The hot-spot temperature rise over oil versus watts per pound shown in Table IV is given in Figure 8. The rises for 30-degree-centigrade oil were corrected by varying the tested rises as the one-fourth power of the viscosity of the average oil film. The slope of the corrected curve is the same ($1/n = 0.7$) as that for ventilated coils, which is reasonable since a large part of the hottest-spot

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*See Appendix II for derivation.

Table IV

Amp Per Square Inch	Watts* Per Pound of Copper	Oil Temp, Deg C	Outside Surface Rise Over Oil, Deg C	Hottest Spot or Maximum Internal Temp Rise Over Outside Surface Temp, Deg C	Hottest Spot or Maximum Internal Temp Rise Over Oil Temp, Deg C
1,000	2.55	22.3	28	22	50
1,740	9.80	39.5	50.5	65	115.5

*At point of maximum temperature.

temperature rise over oil is in the oil film on the outside surface of the coils.

Maximum Temperature Limits for 1-Minute Faults as Governed by Aging of Insulation

Two cases have been worked out—one that cools rapidly, and one that cools slowly—with temperatures to produce the same aging of insulation in 30-degree-centigrade oil. These cases are given in Figure 9. The formula for calculating the aging by the 8-degree-centigrade rule is,

$$\text{Aging} = t\left[\frac{\epsilon^{0.0865T_2} - \epsilon^{0.0865T_1}}{0.0865(T_2 - T_1)}\right] \tag{4}$$

where

t = time in minutes

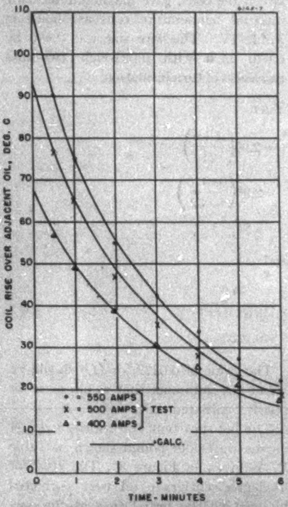


Figure 7. Curves of tested and calculated cooling after 1-minute heating tests

Tests made in 71-degree-centigrade oil
Size of conductor 0.435 by 0.095 inch
Solid line calculated by equation 3

T_2 = maximum temperature

T_1 = minimum at some value below which the aging is negligible (as for example, 100 degrees centigrade in these conditions)

The calculation of the aging for the areas represented by cooling curves *A* and *B* are as follows:

Curve A (fast cooling)

$$\text{Aging} = 2.4\left[\frac{\epsilon^{0.0865\times 202} - \epsilon^{0.0865\times 100}}{0.0865(202-100)}\right]$$

$$= 10.2\times 10^6$$

Curve B (slow cooling)

$$\text{Aging} = 27\left[\frac{\epsilon^{0.0865\times 170} - \epsilon^{0.0865\times 100}}{0.0865(170-100)}\right]$$

$$= 10.7\times 10^6$$

The quantity 10.7 × 10^6 is approximately 1/250 of that resulting for the time (0.0085 weeks or 86 minutes) to reduce the insulation when aged at 200 degrees centigrade to 50 per cent of initial strength as shown in Figure 4 of Clark's paper.[2] That is,

$$\text{Aging} = \frac{86\,\epsilon^{0.0865\times 200}}{250}$$

$$= 10.9\times 10^6$$

In other words, by Clark's data, 250 applications (10 per year for 25 years) would reduce the tensile strength to approximately half its initial value.

Curves *A* and *B* in Figure 9 shows

1. Three-quarters of the aging of the fast cooling windings occurs during the cooling period, while practically 100 per cent of the aging of the slow cooling windings occurs during the cooling period.

2. A difference of approximately 30 degrees centigrade temperature limit should be allowed between a fast cooling and the slowest cooling windings.

3. The temperature limit should be approximately the same for heating periods or faults of one minute or less duration.

AGING CURVES AND TEMPERATURE LIMITS

Figures 10A, B, C, and D show four aging curves, two for an ambient temperature of 30 degrees centigrade and two for an ambient temperature of 75 degrees centigrade, two for 0.019-inch insulation and two for 0.073-inch insulation. These curves show the relative aging on a per unit basis for one event, for varying temperatures, and both for a 10-second and a 1-minute heating time. On a basis of ten events per year, the relative aging per event would be 0.1 and the corresponding temperature limits are as follows:

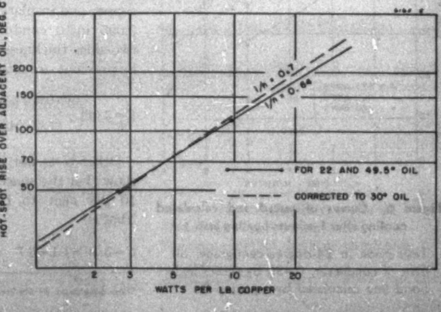
Insulation	Temp Limit, Deg C For 30-Deg-C Oil Temp	Temp Limit, Deg C For 75-Deg-C Oil Temp
0.019	205	201
0.073	196.5	192.5

The coil having 0.019 insulation may be looked upon as a representative of a fast cooling winding so that actual temperature rises of 175 and 125 degrees centigrade may be permitted for 30 and 75 degrees centigrade ambient temperatures respectively.

However, the temperature rise for the 0.073 insulation coils does not represent adequately the performance of a slow cooling coil. In Figure 11 are shown the cooling curves of these four cases with the temperature limit for a relative aging of 0.1 marked by a dot. It is quite evident that the coils both are cooling rapidly.

Figure 8. Ultimate temperature rise of disk coil stack having no oil ducts—slow cooling winding

Conductor size, 0.400 by 0.040 inch, 0.021-inch insulation
Radial build of coils 3.625 inches
Hottest-spot temperature determined by thermocouples between coils



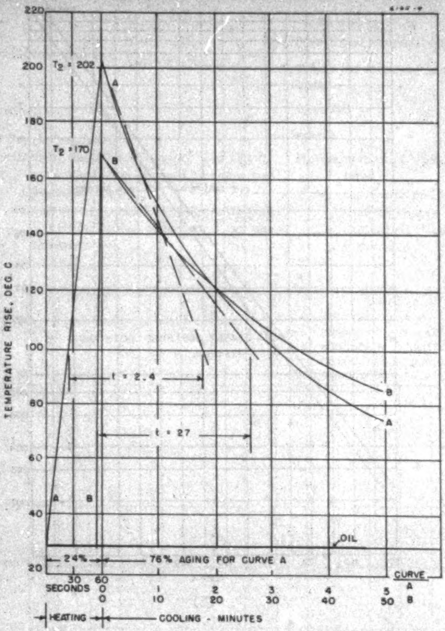


Figure 9. One-minute heating and cooling of windings to produce same amount of aging of insulation

Aging by equation 1

A—Fast cooling winding, 10.2 × 10^8 units

B—Slow cooling winding, 10.7 × 10^8 units

Note:

Heating assumed a straight line

Cooling calculated by equation 3

Initial loss taken from Figure 4 for *A*, and from Figure 8 for *B*

It has been proposed in the current revision of AIEE Standard 32 that coils without ducts be restricted to a rise 25 degrees centigrade less than that obtained for coils with ducts (well-ventilated), or as they may be termed, fast cooling coils. However, this decrement should be taken below the value for the 0.073-inch insulated coils, and the temperature rises then would be 140 and 90 degrees centigrade for ambient temperatures of 30 and 75 degrees centigrade respectively. These permissible temperature rises give ultimate temperatures of 170 and 165 degrees centigrade, slightly higher than those formerly allowed in AIEE Standard 32 when no distinction was made in reference to the rate of cooling.

Basis for Setting Temperature Limits Under Fault Conditions

Temperature limits for windings having any rate of cooling can be based on the initial rate of cooling as determined by the thermal capacity of the windings and the relationship of loss and ultimate rise as shown in the following discussion.

Since practically all the aging occurs during the cooling part of the cycle, it is logical to associate the temperature limit requirements with the rate of cooling. A reasonable criterion for the rate of cooling to determine the suitability of the windings for the temperatures proposed in AIEE Standard 32 is that the time to cool a specified initial part of the cooling range be no greater than a specified amount when calculated by a standardized method.

According to curves *A* and *B*, Figure 9, the time to cool the first ten per cent of the cooling range can be estimated close enough for practical purposes by the following formula:

$$t=k\left(\frac{\Theta_t C}{W}\right) \qquad (5)$$

where

t = time in minutes

k = 0.11 (representing ten per cent cooling + one per cent factor)

Θ_t = initial temperature rise over oil

C = thermal capacity of copper + insulation

W = watts per pound of copper dissipated at a temperature of Θ_t

For example, for curve *A*, Figure 9

where

Θ_t = 172 degrees centigrade (202 − 30)

C = 3.7

W = 225 (Figure 4)

$$t=0.11\left(\frac{172\times3.7}{225}\right)$$

= 0.31 minute, or 18.6 seconds as compared with 18 seconds, curve *A*, to cool 17.2 degrees centigrade

For curve *B* where

Θ_t = 140 (170 − 30)

W = 12 (Figure 8)

C = 3.85

$$t=0.11\left(\frac{140\times3.85}{12}\right)$$

= 4.95 minutes as compared with 4.8 minutes, curve *B*, to cool 14 degrees centigrade

Appendix I. Short-Time Heating Calculations

For a constant current, and when all heat is stored in copper or in copper plus associated insulations, the increase in temperature of copper can be calculated by the following formulas:

Based on Watts Per Pound of Copper

The time t required to produce any temperature rise may be calculated by the following expression:

$$t=\left[\frac{C}{W_0(1-E_0)a_0}\right]\times \frac{1}{2}\log_e\left[(1+a_1\Theta)^2(1-E_0)+E_0\right. \qquad (6)$$

in which

t = time to heat up to a temperature rise of Θ degrees centigrade from a starting temperature of Θ_0

C = thermal capacity of one pound of copper and associated insulations (if C is in watt-seconds t is in seconds, if C is in watt-minutes t is in minutes)

W_0 = total loss in watts per pound at the starting temperature

E_0 = eddy current loss at starting temperature per unit in reference to W_0

a_0 = temperature coefficient of copper at starting temperature

Θ = temperature rise, degrees centigrade

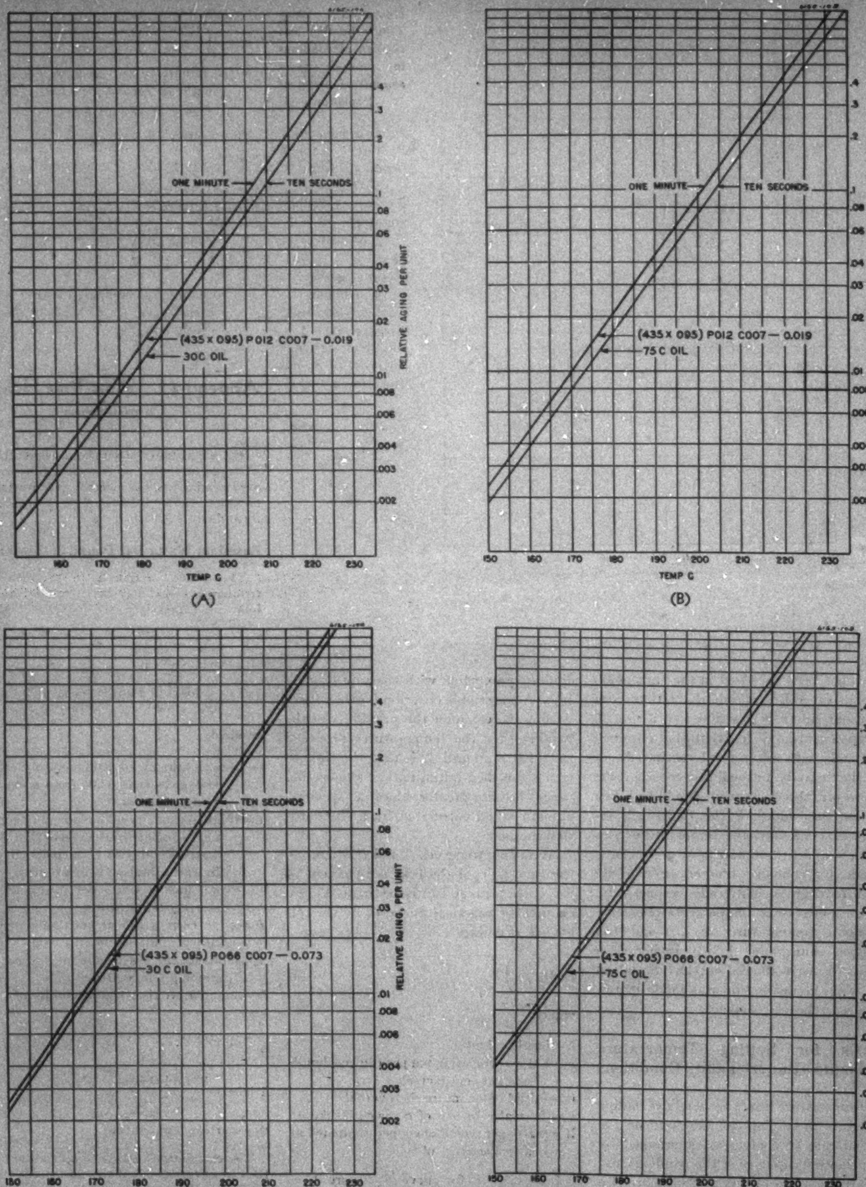
$$\frac{1}{a_0}=234.5+\Theta_0$$

$$1+a_0\Theta=\frac{234.5+\Theta+\Theta_0}{234.5+\Theta_0}$$

Θ_0 = starting temperature

The calculations will be simplified and on the conservative side if the measured loss is used for W_0 and the eddy current loss neglected. The expression then becomes

$$t=(234.5\times\Theta_0)\frac{C}{W_0}m\log\left(\frac{234.5+\Theta_0+\Theta}{234.5+\Theta_0}\right) \qquad (7)$$



(A)

(B)

(C)

(D)

Figure 10. Relative aging curves for two thicknesses of conductor insulation at 30- and 75-degree centigrade oil

Relative aging—one unit = one year at 95 degrees centigrade or

$$\Theta=(234.5+\Theta_0)\left[\log^{-1}\left(\frac{tW_0}{mC(234.5+\Theta_0)}\right)-1\right] \qquad (8)$$

in which

t = time in minutes

C = thermal capacity of copper and associated insulation in watt-minutes per pound of copper

W_0 = watts per pound of copper at starting temperature Θ_0, total based on loss measurement

Θ_0 = starting temperature

Θ = temperature rise at end of time interval t

log = common logarithm

m = conversion factor from common to natural logarithms = 2.3026

Based on Amperes Per Square Inch

When the eddy loss is negligible and the current density is known, the temperature rise can be calculated by

$$\Theta=\log_{10}^{-1}\left[2.00D^2t10^{-11}+\log_{10}(\Theta_0+234.5)\right]-(234.5+\Theta_0) \qquad (9)$$

where

Θ = temperature rise over the starting temperature, Θ_0

D = current density in amperes per square inch

t = time in seconds

Θ_0 = initial temperature, degrees centigrade

When the adjacent insulation absorbs some heat

$$\Theta_c=R_c\Theta$$

where

Θ_c = temperature rise with insulation factored

R_c = ratio of thermal capacity of copper to that of copper plus adjacent insulation when conductor insulation only is factored

$$R_c=\frac{2a}{A+a}$$

where

A = cross section of insulated conductor

a = cross section of bare conductor

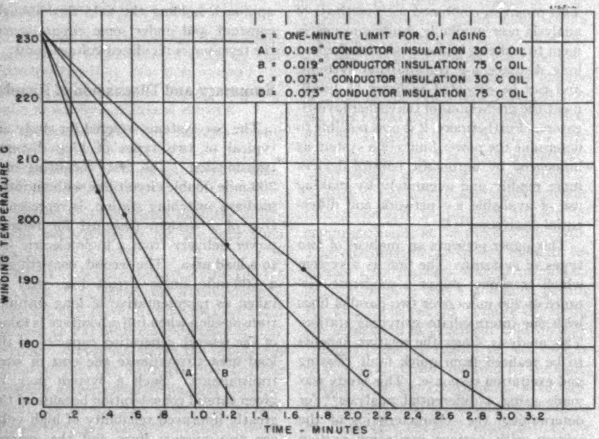


Figure 11. Cooling curves for coils shown in Figure 10

Appendix II. Cooling Calculations

Following is the derivation of formulas for calculating the cooling of a winding where the ultimate rise is not directly proportional to the loss:

C = heat storage capacity of winding, watt-minutes per pound per degree centigrade

K = heat transfer coefficient, winding to oil, watts per pound (per degree centigrade)$^{n'}$

Θ = temperature rise of winding, degrees centigrade

Θ_i = initial temperature rise of winding at start of cooling period

t = time after start of cooling, minutes

n = exponent of ultimate temperature rise Θ versus loss

$K\Theta^n$ = heat loss, winding to oil, watts per pound

Differential equation:

$$K\Theta^n dt=-Cd\Theta$$

$$\frac{d\Theta}{\Theta^n}=-\frac{K}{C}dt \qquad (10)$$

Solution:

$$\frac{\Theta^{1-n}}{(1-n)}=\frac{\Theta_i^{1-n}}{(1-n)}-\frac{K}{C}t$$

$$\left(\frac{\Theta}{\Theta_i}\right)=\left[1+\frac{(n-1)K\Theta_i^n}{C\Theta_i}t\right]^{\frac{-1}{1-n}} \qquad (11)$$

If W_0 = watts per pound to maintain steady-state temperature rise Θ_0, then

$$W_0=K\Theta_i^n$$

and equation 11 becomes

$$\Theta=\Theta_i\left[1+(n-1)\frac{t}{B}\right]^{\frac{-1}{n-1}} \qquad (12)$$

where

$$B=\frac{\Theta_iC}{W_0}$$

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Long-Distance Power Transmission as Influenced by Excitation Systems

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THE renewed attention now being given to transmission of power from hydroelectric sites located at moderate and long distances from their load areas requires a re-evaluation of some of the factors affecting power system stability. This paper shows that the electrical loadings of long distance transmission systems may be increased a significant amount by the use of automatically controlled excitation and proper co-ordination of the apparatus characteristics.

About 20 years ago, particularly during the years 1924, 1925, and 1926, many papers and discussions were presented before the AIEE dealing with the problems of long distance transmission. During this period most of the important factors which influenced the problem, particularly stability, were discussed and methods of analysis described. Much attention[1–7] was given to the use of intermediate synchronous condensers provided with quick acting voltage regulators of the vibrating type.

Since this earlier period, several developments have taken place which make it desirable to reconsider the use of automatically controlled field excitation as a means for increasing the stability power limits of transmission systems. Most important of these has been the development of the continuously acting static type of voltage regulator.[8] During the last war, such voltage regulating means were applied successfully to an important type of naval vessel using electric propulsion.[9] Prior and subsequent to this application, analyses[10] and tests were made which bore out the practicability of using lower short-circuit ratio synchronous machines for this type of drive than would be possible without a specially designed excitation system. Following this experience, tests were made on hydroelectric generators indicating that the static type of continuously acting voltage regulator with a properly co-ordinated excitation system can increase substantially the power limits. This progress, combined with methods of analysis now available, lends encouragement to realizing higher power loadings on long distance a-c transmission systems, and can be expected to bring to fulfillment the predictions of the earlier investigators. Furthermore, it is now possible to determine the power limits of a system as influenced by automatic control devices more readily and accurately by making use of available a-c network and differential analyzers.[11, 12]

This paper presents an analysis of two types of systems. The first is a system which transmits power a moderate distance of 200 miles over two parallel lines with one intermediate switching station. This analysis shows the relative benefits to be realized from quick fault clearing and excitation response. This study was made using a differential analyzer[13] for determining the characteristics of the system with various excitation system parameters. The second system analyzed transmits power over a single-circuit 600-mile transmission line with intermediate synchronous condensers using controlled field excitation, and compares these results with those obtainable without such compensation. This latter part of the study was made with the aid of an a-c network analyzer[14] and is an extension of previous work presented in reference 15. The effect of intermediate synchronous condenser capacity, its location along the line, reactance, and excitation system characteristics are studied. It is shown that with low-transient-reactance synchronous condensers with properly controlled field excitation, an entirely practical method of increasing the power limit becomes available which requires consideration along with other methods of increasing the stability limits. This method of compensation by intermediate synchronous condensers becomes practical because of the ability now to design reliable voltage regulating means which will, in effect for stability purposes, closely approach holding the field flux linkages constant and under some circumstances the terminal or regulated voltage point.

Summary and Discussion of Results

The two systems selected for study are typical of two types of long distance transmission. The first, consisting of a 200-mile double circuit line with an intermediate switching station, is representative of a system designed for reliable power delivery from a hydroelectric site to a load area. The second, consisting of a 600-mile single circuit line, may be taken as representative of long distance transmission when full advantage is taken of the reserve generating capacity of the load area to minimize the cost of such transmission. Such a system may be given careful consideration because of the greatly improved reliability of high voltage transmission lines and the general necessity for reserve generating capacity in the load area.

Although space limitations do not allow a full detailed presentation of the results of these studies, a few selected cases will be used for illustration.

200-Mile Double-Circuit Line

Four typical differential analyzer runs will be described to illustrate the performance of the system of Figure 1 following a 3-phase fault near the generator

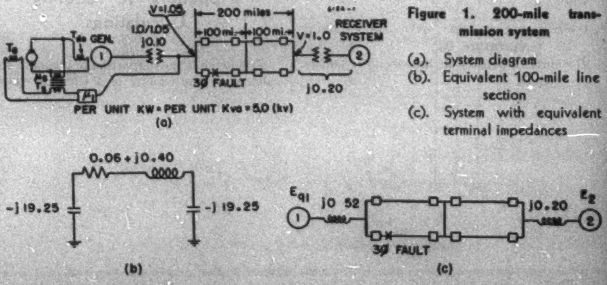

Figure 1. 200-mile transmission system

(a). System diagram
(b). Equivalent 100-mile line section
(c). System with equivalent terminal impedances

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Table I. Summary of Differential Analyzer Runs Used for Illustration

Figure	Run Numbers	P_1	T_e, Sec	t_c, Sec	Remarks
5	48, 76	1.00	∞	0.07	No regulator
6	83	1.05	∞	0	No regulator
7	51, 57	1.10	0.73	0.03	Regulator and stabilizer
8	51, 58	1.10	0.73	0.04	Regulator and stabilizer

high-voltage bus. See discussion under "Analysis" for a description of method and system quantities. The system per unit kilowatt and kilovolt-ampere base is 5.0 × (kv)^2 which for a 230-kv system corresponds to 264,500 kw and kva, and for 360 kv corresponds to 650,000 kw and kva.

For each differential analyzer run the following plots were made:

1. Relative electrical angular displacement between the generator and receiving end system versus relative angular velocity, that is, δ_{12} versus $\frac{d\delta_{12}}{dt}$.
2. Relative electrical angular displacement between the generator and receiving end system versus time, that is, δ_{12} versus t.
3. Generator electrical angular displacement versus time, that is, δ_1 versus t.
4. Generator field flux linkages versus time, that is, e_{q1}' versus t.
5. Generator field current versus time, that is, e_{f1} versus t.

Figures 5 to 8 inclusive show the differential analyzer plots for the four runs summarized in Table I.

Figure 5 as indicated by Table I is for the case of the generating station initially carrying unit power with no regulator and having a total fault clearing time of 0.07 second. Figure 5A shows the oscillation between the per unit relative angular velocity and the relative angular displacement. As shown, the system is initially stable and has a damped oscillation which is decaying to about 90 degrees. From this plot stability may or may not be maintained, depending upon whether or not the field flux linkages will be reduced further. Figures 5B and 5C show the oscillation of δ_{12} versus time and δ_1 versus time. As shown by Figure 5B, the relative angular displacement indicates a damped oscillation, but the average value gradually is increasing. Figure 5C indicates the decrease in speed which results from the increased losses in the transmission line and which has not been compensated for by governor action, as change in prime mover torque was neglected during this initial period of the disturbance. Figure 5D shows the decay in field flux linkages. Initially, the decay is relatively rapid because of the fault. Following the clearing of the fault, the linkages continue to decrease and, as indicated, still are decaying gradually at the end of 5.5 seconds. This continual decay of flux linkages indicates that stability will be ultimately lost although it was maintained during the first three swings. Figure 5E shows the oscillation in generator field current. Initially there is a rise in the d-c component of field current induced by the short circuit (the a-c components were neglected), and then an oscillation following the clearing of the fault resulting from the rotor oscillation. As shown, the average field current is above the initial value indicating that there is a component of induced current resulting from the gradual increase in relative angular displacement of the rotor. All of these results confirm the indication that stability cannot be maintained even for this loading if no regulator is used. Other runs show that with a regulator, even if the excitation response is very low, stability can be maintained for such a case as Figure 5.

Figure 6 is similar to Figure 5 in that no regulator is used. The loading has been increased to 1.05 per unit. The fault clearing time is zero corresponding to "no fault" or merely opening a generator-end transmission-line section. As indicated by Figure 6A, although stability is maintained during the first three swings of the oscillation, it is subsequently lost because the angular displacement after the third swing increases indefinitely. The curves of Figure 6 can be compared with those of Figure 5 to show the effect of a slightly higher initial loading with a reduced switching time. As shown by Figure 6D the field flux linkages show a definite trend downward. Figure 6E shows the induced field current resulting from the oscillation and the displacement of the average current from the initial value resulting from the gradual increase in average relative angular displacement.

It is evident from a comparison of Figures 5 and 6 that the conditions of Figure 5 are less unstable but may be considered to be very near the critical load when no regulator is used. Both of these cases, as is shown by other runs, are stable even with a relatively slow excitation system response.

Figure 7 is for the case of higher loading ($P = 1.1$) than for the cases of Figures 5 and 6. For $P = 1.1$, a regulator using an excitation system stabilizer is used and the excitation system time constant is of moderate value ($T_e = 0.73$ second). This corresponds to a per unit excitation system response of approximately 1.0. During the first swing and subsequently stability is maintained, the oscillation being damped out toward a final steady-state value. This is borne out by the other curves of Figure 7, with the field flux link-

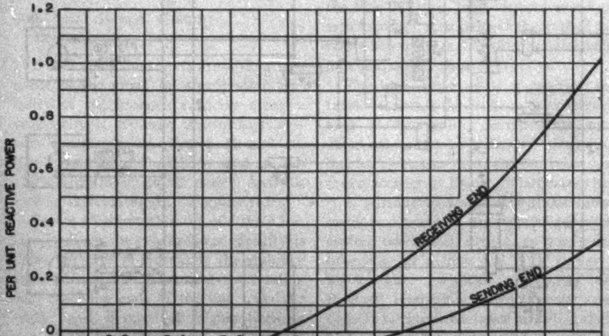


Figure 2. Transmission line real power and reactive power circle diagram for all line sections of both lines in operation

System of Figure 1

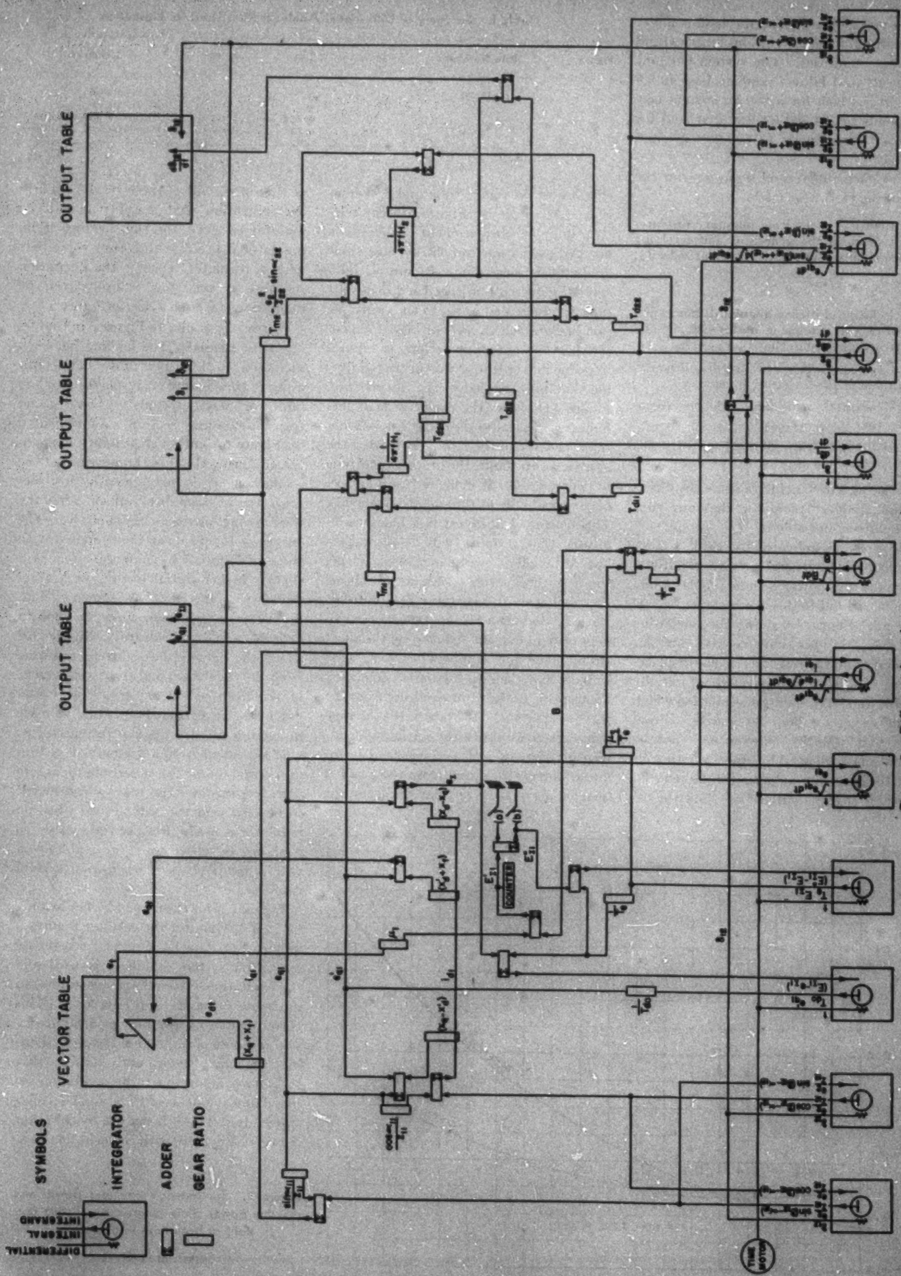

SYMBOLS
INTEGRATOR
ADDER
GEAR RATIO
DIFFERENTIAL
INTEGRAL
INTEGRAND
VECTOR TABLE
OUTPUT TABLE
OUTPUT TABLE
OUTPUT TABLE
COUNTER
TIME MOTOR

Figure 3. Differential analyzer connection diagram

ages and field current settling out to new sustained values.

Figure 8 is for a case similar to Figure 7 except that the switching time has been increased from 0.03 to 0.04 second. Stability is lost on the first swing although the field flux linkages actually are increased following the fault period resulting from the response of the excitation system.

From the results of many runs similar to these Figure 9 was obtained which shows the change in transient power limit with fault clearing time and with excitation system response. It is interesting to note that for a practical switching time for the assumed system, that is, $t_s \approx 0.08$ second or 4.8 cycles, the excitation system response does not influence appreciably the maximum power that can be carried. However, with more rapid fault clearing times, the effect of the excitation system response is more pronounced. These results may be interpreted for less severe faults by assuming the 3-phase fault to be of shorter duration than the actual switching time. That is, for a line-to-ground fault actually switched in 4.8 cycles, it would be expected that the resulting performance would be essentially not much different from that obtained with zero switching time, as the severity of a line-to-ground fault is relatively small. Also, for faults located at a distance from the high-voltage bus, the fault severity would be decreased correspondingly. The results can be interpreted therefore to indicate that greater benefits can be expected by the more rapid response excitation systems for faults of reduced severity than for very severe faults.

600-Mile Single Circuit Transmission

The most significant results of the study made of the effect on steady-state stability of synchronous capacity along a single circuit 600-mile transmission system are shown on Figures 15 and 16. The constants and methods used are described in more detail under Analysis and by information in the figure titles.

Figure 15 shows that for relatively low line loadings the synchronous capacity is more effective in increasing the stability limit if it is concentrated at the sending and receiving ends of the line. On the other hand, if relatively higher loadings are to be used, the same synchronous capacity is more effective if it is distributed between the sending and receiving end and the intermediate stations. Although Figure 15 illustrates the relative advantages of distributing capacity along the line to attain high transmission system loadings, it is only of theoretical interest because it is difficult to realize such an equal distribution in practice.

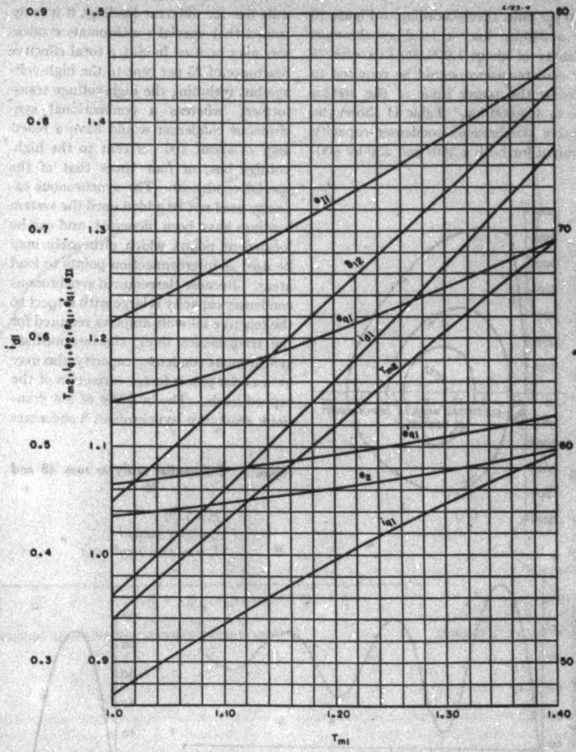


Figure 4. Initial quantities before fault

Figure 16, however, is of practical interest because it shows the relative advantage of distributing intermediate synchronous condenser capacity along the line for given generator and receiving system reactances. Figure 16 indicates that it is possible to transmit without intermediate condensers and typical terminal system reactances, a receiver power approaching as a limit 1.9 (kv)^2 or for a 230-kv line, $1.9 \times 230^2 = 100{,}000$ kw. If three intermediate synchronous condensers having a reactance including their high voltage transformer of 25 per cent on their rating with a total rating of $2.5 \times 230^2 = 132{,}000$ kva are placed along a line, their total per unit admittance is 4.0 and the steady-state power limit is increased to $2.85 \times 230^2 = 151{,}000$ kw. Such a loading is quite reasonable for a 230-kv line. If the line is operated at 360 kv, the corresponding power limit would be 369,000 kw with an intermediate synchronous capacity of 324,000 kva. Such an amount of intermediate synchronous capacity is entirely reasonable and practical.

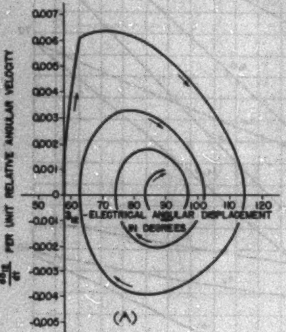
The practicability of this method depends upon obtaining a low effective reactance as a result of being able to use a voltage regulator which in effect will hold the voltage behind transient reactance of specially designed low transient reactance synchronous condensers. If, however, it were not possible to build such low transient reactance synchronous condensers, or if it were not possible to depend on the voltage regulator characteristics, the required synchronous condenser capacity would be increased three to four times, making such a method uneconomical in comparison with other methods of line compensation.

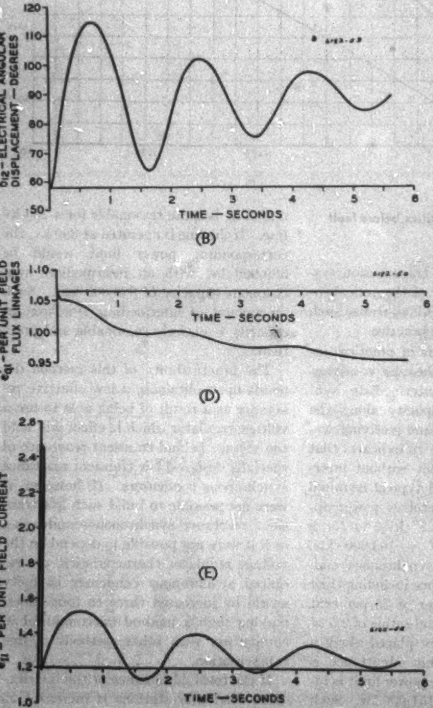
If the total admittance of the intermediate condenser stations is increased to

12.0 per unit, then according to Figure 16 for a 360-kv line, a total synchronous capacity of about 1,000,000 kva with 25 per cent reactance would be required to increase the power limit of the 360-kv line to 475,000 kw. Table II shows the relative synchronous condenser capacity required for both a 360- and 230-kv 600-mile line for different loadings if it is assumed that special synchronous condensers may be used having a total effective reactance of 25 per cent to the high-voltage bus including the high-voltage transformer; whereas a conventional synchronous condenser would have a reactance of about 100 per cent to the high-voltage bus, or four times that of the special condenser. The synchronous capacity need not be added until the system loadings have been increased, and can be located at points which ultimately may be used as interconnection points to load areas. Because the required synchronous condenser capacity is large with respect to the reactive kilovolt-amperes required for the transmission lines, the intermediate synchronous condenser capacity also may be used for power factor correction of the tapped loads. The building of low transient reactance synchronous condensers with special excitation systems now appears entirely practical and makes it necessary to consider such a method in the design of any new long distance transmission system. However, if it were not possible to attain the special features, the required synchronous condenser capacity would become extremely large, making this method uneconomical compared with other methods.

A 600-mile line has an appreciable line charging capacity, approximately 166,000 kva for a 230-kv line and 405,000 kva for a 360-kv line. The intermediate synchronous condenser capacity also may provide the required underexcited capacity during periods of light load and thereby maintain normal voltage levels along the line.

It also is evident from Figure 16 that it may not be desirable economically to locate the synchronous condensers along the line in any more than two or three locations because of the diminishing gain realized by increasing the number of intermediate stations. Because of the relatively large required capacity, however, it becomes economically possible to locate the synchronous condensers at more than one location. A margin in the steady-state stability under automatic control may be taken so as to allow for failure of any one condenser or its excitation system and also to allow the system to ride through transient faults in the terminal systems. The required margin will depend on the requirements of the particu-





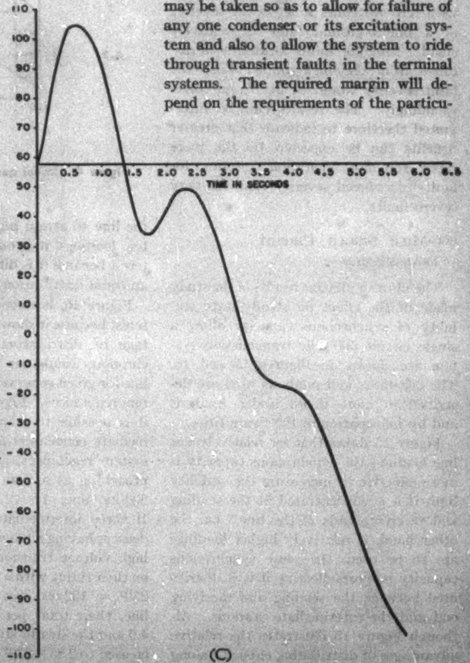


Figure 5. Differential analyzer runs 48 and 76

$P_1 = 1.00$
$t_1 = 0.07$ second
$T_e = \infty$ (no regulator)

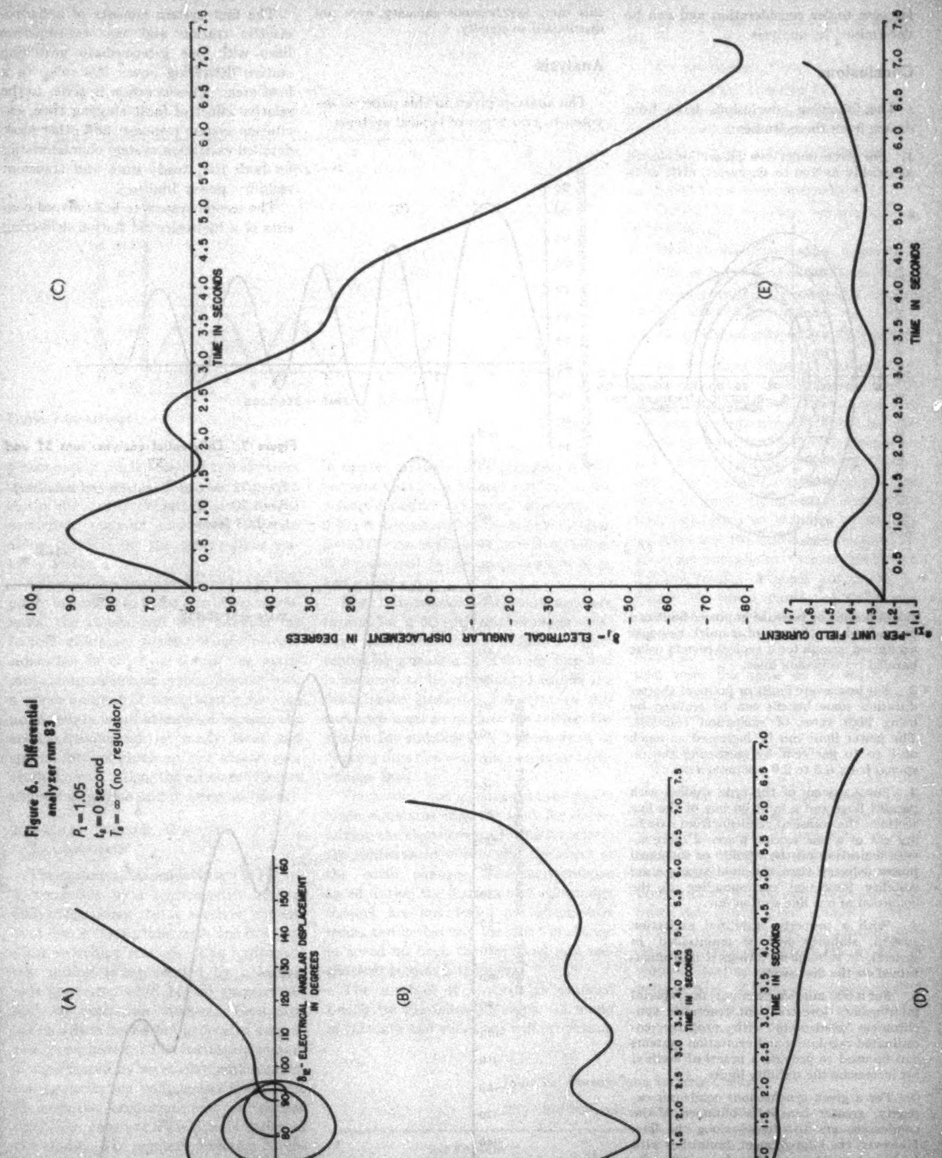


Figure 6. Differential analyzer run 83

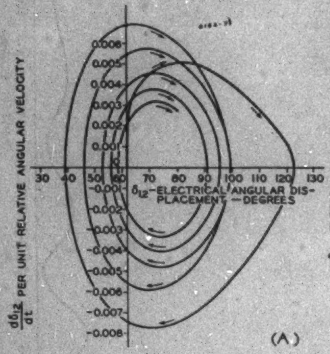
$P_1 = 1.05$
$t_c = 0$ second
$T_e = \infty$ (no regulator)

lar case under consideration and can be determined by analysis.

Conclusions

The following conclusions have been drawn from these studies:

1. For severe faults on a line section cleared as quickly as five to six cycles, little additional benefit or increase in power limit can be realized by the use of a quick response excitation system for a typical system using parallel transmission lines.



2. For less severe faults or faults of shorter duration some benefit can be realized by using high rates of excitation response. The power limit can be increased as much as 5 to 10 per cent by increasing the response from 0.5 to 2.0 per unit.

3. For a system of the type studied with parallel lines and a fault on one of the line sections, the transient resulting from switching out of a line section places a more severe limitation on the reliable or sustained power delivery than the final steady-state stability limitation corresponding to the condition of one line section out.

4. With a properly stabilized excitation system, stability will be maintained, in general, on subsequent swings if it is maintained on the first swing.

5. For a 600-mile single circuit line, special intermediate low transient reactance synchronous condensers with properly coordinated regulators and excitation systems can be used to provide a practical method for increasing the stability limits.

6. For a given synchronous condenser capacity, greater benefit is obtained if the condensers are distributed along the line. However, the improvement diminishes with increased number so that for a 600-mile line, the optimum economic number probably will be two or three.

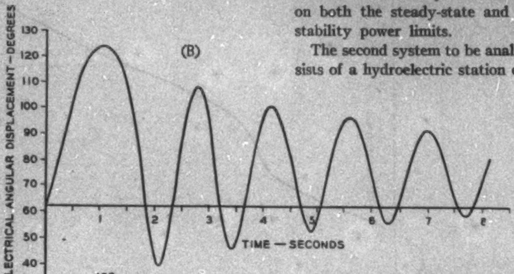
7. A more nearly equal distribution of synchronous capacity along and at the terminals of a long distance transmission system allows for higher loadings than if this same synchronous capacity were not distributed so equally.

Analysis

The analysis given in this paper is devoted to two types of typical systems.

The first system consists of a hydroelectric station and two transmission lines with one intermediate switching station delivering power 200 miles to a load area. Consideration is given to the relative effect of fault clearing time, excitation system response, and other more detailed excitation system characteristics on both the steady-state and transient stability power limits.

The second system to be analyzed consists of a hydroelectric station delivering



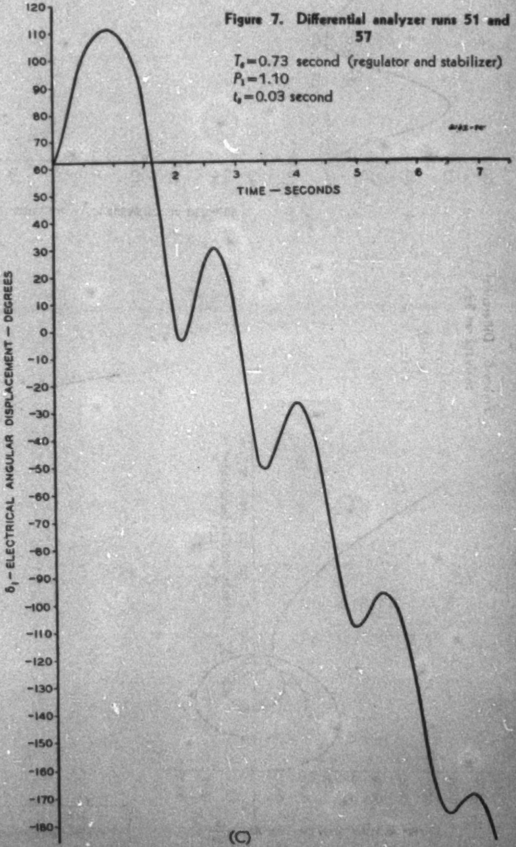


Figure 7. Differential analyzer runs 51 and 57

T_e=0.73 second (regulator and stabilizer)
P_1=1.10
t_c=0.03 second

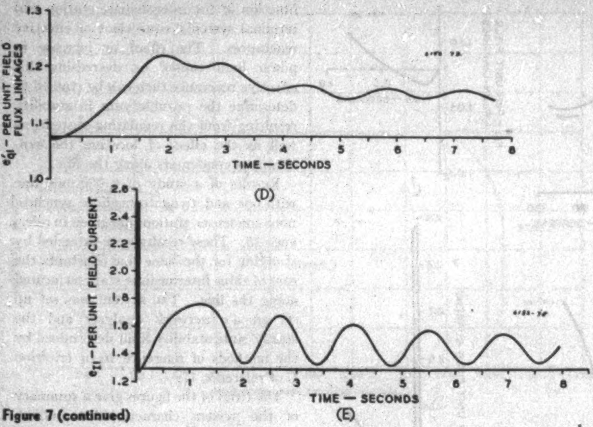


Figure 7 (continued)

power over a single 600-mile transmission line to a load area. Consideration is given to the effect of intermediate synchronous condenser stations and their location along the line on the steady-state stability limits.

Although the analysis presented in this paper is limited to these two types of systems, the results will help indicate the factors which are worthy of careful consideration in order to obtain the maximum transmission or system loading with a given amount of equipment. An economic study is not attempted because the economics depend on many local and special factors which are not always generally known. Also, the economic factors change with time and progress in the art.

200-Mile Double Circuit Transmission

The system analyzed is shown in Figure 1, consisting of a hydroelectric station delivering power to a receiver system over two 200-mile lines with one intermediate switching station. The hydroelectric station is represented by a salient pole generator with step-up transformer and an excitation system. Generator saliency is included but generator saturation is neglected. The excitation system is represented by an exciter with a given ceiling excitation voltage and lower limit of negative excitation and an exciter field time constant T_e which is varied in the study. The regulating system is represented by a regulator amplification factor of μ. This amplification factor is the ratio of the increase in per unit exciter voltage to the per unit decrease in voltage applied to the voltage sensitive element of the regulator to produce this increase in exciter voltage. For example, if 0.01 per unit change in voltage applied to the voltage-sensitive element produces a 0.10 per unit change in the exciter voltage, $\mu = 10.0$. An excitation system stabilizer is represented by an amplification of μ, and a time constant of T_s.

The transmission line constants are typical for a 60-cycle high-voltage overhead line. The receiver system is represented by a reactance of 20 per cent and is assumed to be unregulated during the disturbance studied. The type of disturbance used as a basis for testing the system for stability is a 3-phase fault of varying duration near the generator high-voltage bus.

The assumption is made that the power angle equations may be used for determining the electric torque, which neglects the subtransient effects and the effect of the small changes from synchronous speed during the disturbance. Damping torques are introduced as independent terms and in this way the effect of change in speed on both the electrical and mechanical torque is included.

The method of analysis is outlined briefly by the following steps. A study of the data and equations will provide an understanding of the basis for the analysis.

1. System diagram, Figure 1.
2. System data, Appendix II.
3. Transmission line circle diagram (both lines, all sections in), Figure 2.
4. Characteristic system impedances, Appendix III.
5. System equations, Appendix IV.
6. Differential analyzer equations, Appendix V.
7. Different analyzer diagram, Figure 3.
8. Plot of initial conditions, Figure 4.
9. Typical differential analyzer output curves, Figures 5, 6, 7, and 8.
10. Graphical summary of results, Figure 9.

The system of Figure 1 was analyzed also for the steady-state power limit with one of the sending end transmission line sections opened, corresponding to the condition after the fault was removed. The same system equations (Appendix IV) were used to obtain relations for the condition of small variations. The system then was tested for stability by Routh's criterion and the stable and unstable regions determined for various conditions. Typical results obtained are shown by Figure 10, which shows the stable and unstable regions of generator power as a function of the regulator amplification factor μ. The system parameters for this case were the same as indicated previously except as noted in the caption of Figure 10.

600-Mile Single Circuit Transmission

Analyses of steady-state stability similar to those presented in reference 10 and in Figures 10 and 11 indicate that it is possible to operate at appreciably higher power loadings under steady-state conditions with excitation systems designed to raise the steady-state stability limits. These results indicate, in general, two limits or levels, one which readily can be realized corresponding to holding the voltage back of transient reactance, and one which is more difficult to realize corresponding to holding the regulated

Table II. Steady-State Stability Limits

600-Mile 60-Cycle Single Circuit System

230 Kilovolts			360 Kilovolts		
Power Received, Kw	Synchronous Condenser Kva		Power Received, Kw	Synchronous Condenser Kva	
	Special	Conventional		Special	Conventional
101,000	0	0	247,000	0	0
156,000	132,000	528,000	383,000	324,000	1,296,000
195,000	264,000	1,056,000	475,000	648,000	2,592,000

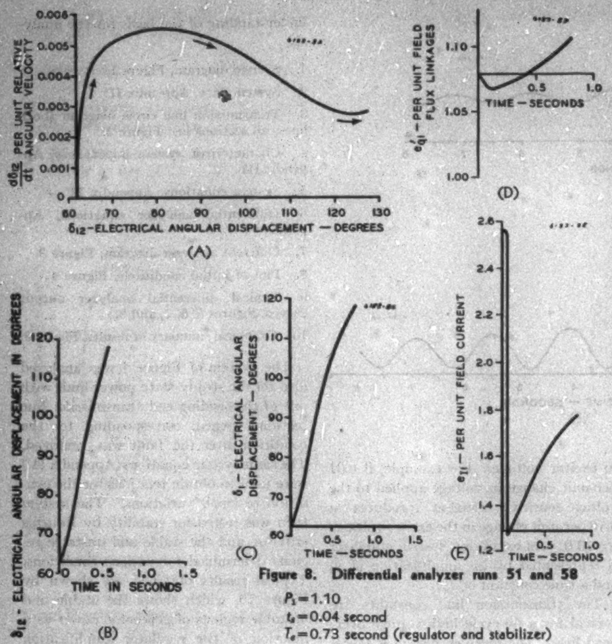


Figure 8. Differential analyzer runs 51 and 58

$P_1 = 1.10$
$t_s = 0.04$ second
$T_e = 0.73$ second (regulator and stabilizer)

voltage constant. The results shown by Figure 11 obtained for the analysis in Appendix I indicate that an intermediate condenser station may be used to raise the limit of the transmission system to a level corresponding to holding approximately the intermediate synchronous condenser's voltage back of transient reactance or with high amplification factors approximately the regulated voltage point in the system. This suggests the desirability of studying the power limits of a long line with intermediate synchronous condensers as a function of the intermediate station and terminal systems' equivalent or effective reactances. The effect or increase in power limit caused by decreasing this effective reactance then can be studied to determine the possible gain in stability resulting from the regulating systems as well as the effect of locating the synchronous condensers along the line.

Results of a study of a 600-mile line with one and two intermediate synchronous condenser stations are given in reference 15. These results were extended by studying for the same line constants the case of three intermediate stations located along the line. The system was set up on an a-c network analyzer and the steady-state stability limit determined by the methods of reference 16 or criterion *C* of reference 17.

The titles of the figures give a summary of the system characteristics studied. Figure 12 gives the power angle characteristic, where δ_{12} is the angular displacement between the sending and receiving terminals of the line. The required reactive power for the intermediate synchronous condensers and the line terminals is shown also. The vertical lines indicate the steady-state stability limit for the indicated values of station reactances when $X_G = X_M = X_{c1} = X_{c2} = X_{c3}$. This shows the change in power limit and reactive

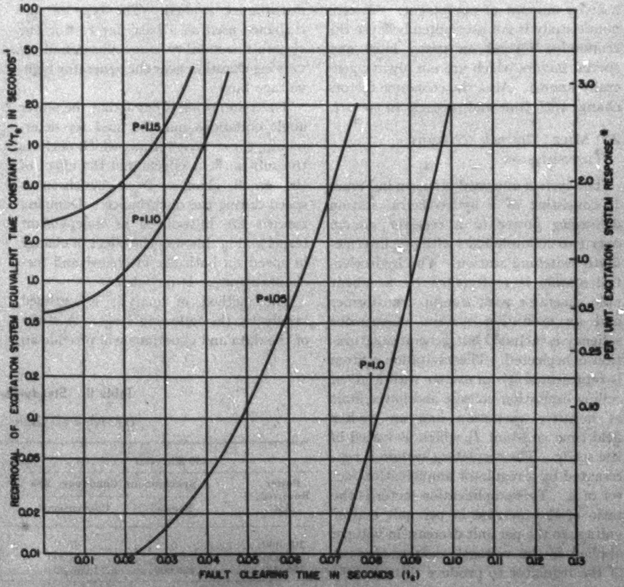
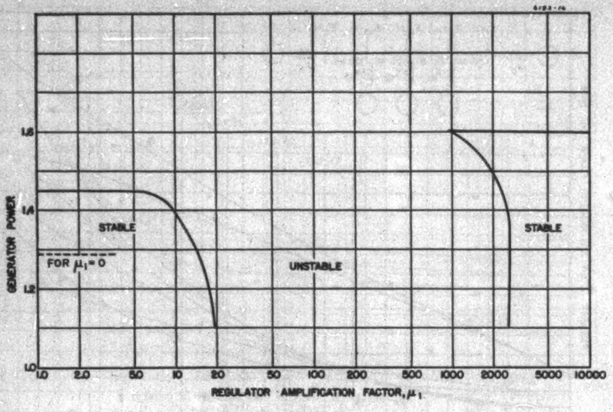


Figure 9. Summary of portion of differential analyzer results showing relation between transient power limit P_1, fault clearing time t_s, and excitation system response for 200-mile transmission system for Figure 1

These results are based on drawing curve of critical switching time for constant initial power between stable and unstable differential analyzer run points

Per unit excitation system response as plotted on right hand scale is the average rate of increase of exciter voltage during the first half second following a sudden drop in a-c regulated voltage expressed as a ratio of exciter volt-seconds and initial exciter voltage. For this scale, initial exciter voltage was taken corresponding to an initial load of $P_1 = 1.0$ and $e_{f1} = 1.217$. For an initial load of $P_1 = 1.15$ rather than 1.0 the per unit excitation response decreases only slightly, for example, from a response of 1.0 to 0.91 per unit for $1/T_e = 1.5$



power requirement with changes in station effective reactance.

Figure 13 is similar to Figure 12 except that the sending and receiving system reactances are kept constant and the abscissa is the total angular displacement δ_{GM}. The vertical lines indicate the power limit and reactance power requirements with change in the intermediate station terminal reactance. Figure 14 is a summary of the results obtained from reference 15 and Figure 13. Figure 14 shows the change in power limit with change in intermediate synchronous condenser reactance. Figure 15 is a summary of the results obtained from reference 15 and Figure 12 with total station admittance as abscissa. This abscissa indicates the total synchronous machine capacity required if it were distributed equally between the terminal and the intermediate stations. This indicates that for higher loadings the synchronous capacity is more effective in raising the stability limit if it is distributed more equally along the line. However, for lower loadings it is more effective if it is concentrated at the ends. Figure 16 is a summary of results from Figure 14. This figure indicates that for given sending and receiving end reactances, the power limit is increased with increase in the number of intermediate condenser stations along the line for the same total synchronous condenser capacities.

Nomenclature

The nomenclature used here follows that which has been established generally by previous usage. All machine and system quantities are in per unit of the designated bases unless otherwise specified.

Figure 10. Steady-state power limits of system of Figure 1 with one generator line section out corresponding to system condition following clearing of fault, showing effect of generator voltage regulator amplification factor

System data same as shown in Appendix II with following exceptions:

$T_{d11} = T_{d12} = T_{d21} = T_{d22} = 3.0$
$T_s = 200$ radians or 0.53 second
$\mu_1 =$ variable

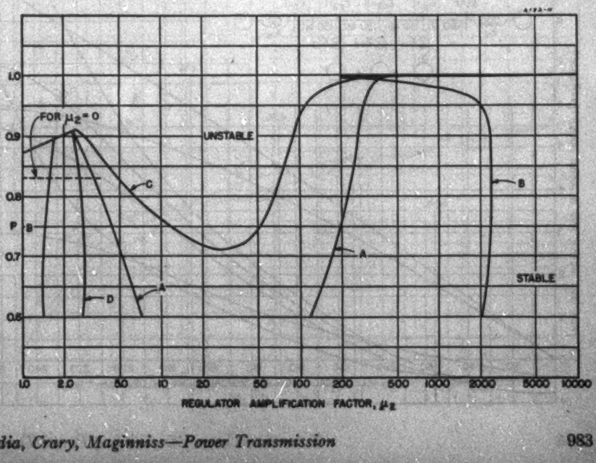
Figure 11. Steady-state power limits of 3-machine stability system of Figure 17

Data as given in Appendix I

A—T_{d1} 0, $T_{d2} = 3.0$
B—$T_{d1} = 3$, $T_{d2} = 0$
C—$T_{d1} = 3$, $T_{d2} = 3.0$
D—$T_{d1} = 0$, $T_{d2} = 0$

D has only one stable region above $P = 0.6$

e = voltage
e_{dt} = direct-axis component of regulated voltage
e_I = quadrature-axis voltage proportional to field current
e_q' = quadrature-axis voltage proportional to field flux linkages
e_{q1} = quadrature-axis voltage of generator back of $x_{q1}i_1$
e_{qt} = quadrature-axis component of regulated voltage
e_t = regulated voltage
E_I = quadrature-axis voltage corresponding to field voltage
f = frequency in cycles per second
H = inertia constant in kilowatt-seconds per kilovolt-ampere
i = armature current
i_d = direct-axis component of armature current
i_q = quadrature-axis component of armature current
$p = d/dt$ = differential operator
t = time in electrical radians or seconds
T = torque
T_d = damping torque
T_{do} = generator open-circuit time constant in radians or seconds
T_e = electric torque
T_m = mechanical torque
T_s = synchronizing torque or excitation system stabilizer time constant
x_d = direct-axis synchronous reactance
x_d' = direct-axis transient reactance
x_q = quadrature-axis synchronous reactance
x_t = transformer reactance
Z_{nm} = transfer impedance
Z_{nn} = driving point impedance
$\alpha = \pi/2 - \theta$
δ = electrical angular displacement in radians or degrees
θ = impedance angle
μ = regulator amplification factor
μ_s = excitation system stabilizer amplification factor



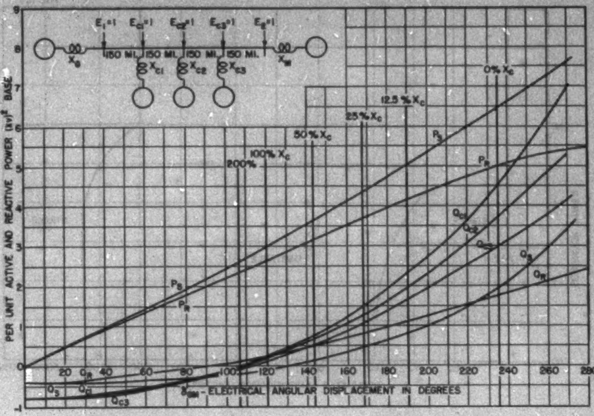


Figure 13. Characteristics of 600-mile 60-cycle transmission line with synchronous terminal equipment and three intermediate synchronous condensers

Per unit kva = per unit kw = (kv)², $x=0.8$ ohm per mile, $r=0.08$ ohm per mile, $b=5.2\times10^{-6}$ mho per mile, $X_G=0.40$ per unit based on 2.5 (kv)² (0.16 per unit based on (kv)²). $X_M=0.25$ per unit based on 2.5 (kv)² (0.10 per unit based on (kv)²) $X_{c1}=X_{c2}=X_{c3}$ and $E_1=E_2=E_{c1}=E_{c2}=E_{c3}=1.0$. Vertical lines give steady-state stability limits with intermediate station reactances based on kva = 2.5 (kv)²

General Subscripts

d = direct-axis

q = quadrature-axis

t = regulated voltage point

1 = generator or sending end system

2 = intermediate synchronous condenser or receiving end system

0 = initial value

Appendix I. Intermediate Synchronous Condenser With Automatic Voltage Regulator

The following analysis is made to determine the effect of a voltage-regulated intermediate synchronous condenser on the steady-state stability limits of a simple type of system. Previous analyses have indicated the gains to be realized for the simple 2-machine case, whereas this analysis is for a 3-machine case.

The system to be analyzed is shown in Figure 17, only machine 2 is assumed to have transient saliency ($x_d'/x_q \neq 1.0$) and only machine 2 has controlled excitation.

The equations for the system, in accordance with the usual assumptions that the electric torque is expressible by the steady-state equation for power flow, are as follows:

$$M_1\frac{d^2\delta_1}{dt^2}+T_{d1}\frac{d\delta_1}{dt}=T_{m1}-T_{e1} \tag{1}$$

$$M_2\frac{d^2\delta_2}{dt^2}+T_{d2}\frac{d\delta_2}{dt}=T_{m2}-T_{e2} \tag{2}$$

where

$M=4\pi fH$

t = time in electrical radians at normal frequency, and

$$T_{e1}=\frac{e_1e_{q2}}{x_{12}}+\frac{e_1e_3}{x_{13}}\sin\delta_{13} \tag{3}$$

$$T_{e2}=\frac{e_{q2}e_1}{x_{12}}\sin\delta_{21}+\frac{e_{q2}e_3}{x_{23}}\sin\delta_{23} \tag{4}$$

$$pe_{q2}'=\frac{1}{T_{do2}}(E_{I2}-e_{I2}) \tag{5}$$

$$e_{q2}'=e_{q2}-i_{d2}(x_{q2}-x_{d2}') \tag{6}$$

$$e_{I2}=e_{q2}+i_{d2}(x_{d2}-x_{q2}) \tag{7}$$

$$E_{I2}-E_{I20}=-\mu_2(e_t-e_{t0}) \tag{8}$$

$$e_{dt2}=-i_{q2}x_{q2} \tag{9}$$

$$e_{qt2}=e_{q2}-i_{d2}x_{q2} \tag{10}$$

$$i_{d2}=\frac{e_{q2}}{x_{22}}-\frac{e_1}{x_{12}}\cos\delta_{21}-\frac{e_3}{x_{23}}\cos\delta_{23} \tag{11}$$

$$i_{q2}=\frac{e_{q2}}{x_{12}}\sin\delta_{21}+\frac{e_3}{x_{23}}\sin\delta_{23} \tag{12}$$

Equations 1 to 12 were solved for the condition of small oscillations and the resulting characteristic denominator tested for stability by Routh's criterion.

Figure 11 is representative of the results obtained for the system of Figure 17 and the following initial conditions.

Constants

$e_1=1.0$

$e_3=1.0$

$x_{12}=3.0$

$x_{23}=3.0$

$x_{13}=3.0$

$x_{22}=1.5$

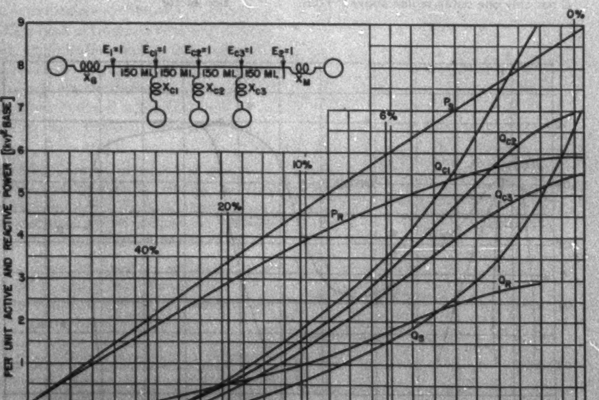
$M_1=4{,}000$

Figure 12. Characteristics of 600-mile conventional 60-cycle transmission line with synchronous terminal equipment and three intermediate synchronous condenser stations

Per unit kva = per unit kw = (kv)², $x=0.8$ ohm per mile, $r=0.08$ ohm per mile, $b=5.2\times10^{-6}$ mho per mile. Vertical lines give steady-state stability limits with equal terminal and intermediate station reactance as indicated based on (kv)²

$X_G=X_M=X_{c1}=X_{c2}=X_{c3}$

$E_1=E_2=E_{c1}=E_{c2}=E_{c3}=1.0$



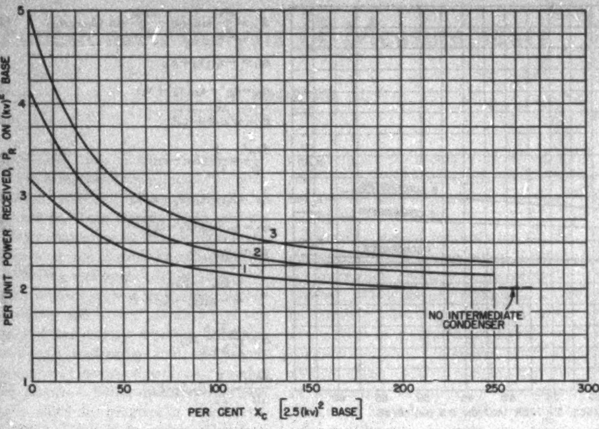


Figure 14. Receiving end power at steady-state limit for 600-mile 60-cycle conventional line with one, two, and three intermediate synchronous condenser stations equally spaced along the line as affected by intermediate station reactances

$X_G=40$ per cent [2.5 $(kv)^2$ base]
$X_M=25$ per cent [2.5 $(kv)^2$ base]
$x=0.8$ ohm per mile
$r=0.08$ ohm per mile
$b=5.2\times10^{-6}$ mho per mile
1—One intermediate condenser station
Line in two 300-mile sections
$E_1=E_2=E_{c1}=1.0$
2—Two intermediate condenser stations
Line in three 200-mile sections
$E_1=E_2=E_{c1}=E_{c2}=1.0$
$X_{c1}=X_{c2}$
3—Three intermediate condenser stations
Line in four 150-mile sections
$E_1=E_2=E_{c1}=E_{c2}=E_{c3}=1.0$
$X_{c1}=X_{c2}=X_{c3}$

$M_2=1{,}500$
$x_{q2}=1.0$
$x_d{}'_2=0.30$
$T_{do2}=2{,}000$
$e_{f2}=1.0$
$e_{qf2}=1.0$
$e_{df2}=0$
$T_{d1}=T_{d2}=0$ (curve *D*, Figure 11)
$T_{d1}=3.0$, $T_{d2}=0$ (curve *B*, Figure 11)
$T_{d1}=3.0$, $T_{d2}=3.0$ (curve *C*, Figure 11)
$T_{d1}=0$, $T_{d2}=3.0$ (curve *A*, Figure 11)
$\sin\delta_{12}=\sin\delta_{23}=P$
$\cos\delta_{12}=\cos\delta_{23}=\sqrt{1-P^2}$
$\cos\delta_{13}=1-2P^2$
$e_2=1+2\left(1-\sqrt{1-P^2}\right)$

Appendix II. System Data

Generator Data

Rated power factor = 0.95
Kilowatt rating = 5.0 × $(kv)^2$
Kilovolt-ampere rating = 5.27 × $(kv)^2$

$$\left.\begin{array}{l} x_d=0.60 \\ x_q=0.40 \\ x_d'=0.20 \end{array}\right\} \text{on } 5.27\times(kv)^2 \text{ or generator kilovolt-ampere base}$$

$$\left.\begin{array}{l} x_d=0.63 \\ x_q=0.42 \\ x_d'=0.21 \end{array}\right\} \text{on high voltage and system base of } 5.0\times(kv)^2\ \left(\frac{x_{gen}\times0.95\times}{1.05^2}\right)$$

$T_{do}=1{,}885$ radians or 5.0 seconds
$H=5.0$ kilowatt-seconds per kilovolt-ampere

Line Data

$x=0.8$ ohm per mile
$r=0.12$ ohm per mile
$y=5.2$ micromhos per mile

System Data (On System Base)

$H_2=50.0$ kilowatt-seconds per kva

$$\left.\begin{array}{l} T_{d11}=4.0 \\ T_{d12}=3.0 \\ T_{d21}=3.0 \\ T_{d22}=18.0 \end{array}\right\} \text{damping torque coefficients}$$

Excitation System Data (Per Unit on System Voltage Base)

Ceiling excitation voltage, $E_{I(max)}''=2.25$
Minimum excitation voltage limit,
$E_{I(min)}''=-0.304$

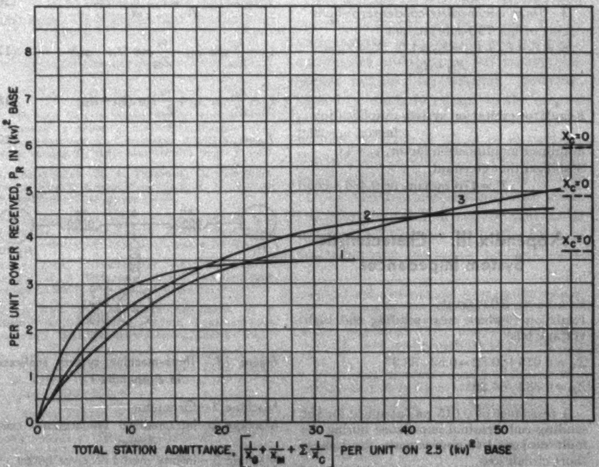


Figure 15. Receiving end power at steady-state limit for 600-mile 60-cycle conventional line with one, two, and three intermediate synchronous condenser stations, equally spaced along the line as affected by total synchronous capacity of terminal and intermediate stations expressed as the summation of the station equivalent per unit admittances on 2.5 $(kv)^2$ base

$x=0.8$ ohm per mile
$r=0.08$ ohm per mile
$b=5.2\times10^{-6}$ mho per mile
$X_G=X_M=X_C$
1—One intermediate synchronous condenser station
Line in two 300-mile sections
$E_1=E_2=E_{c1}=1.0$
$X_c=X_{c1}$
2—Two intermediate synchronous condenser stations
Line in three 200-mile sections
$E_1=E_2=E_{c1}=E_{c2}=1.0$
$X_c=X_{c1}=X_{c2}$
3—Three intermediate synchronous condenser stations
Line in four 150-mile sections
$E_1=E_2=E_{c1}=E_{c2}=E_{c3}=1.0$
$X_c=X_{c1}=X_{c2}=X_{c3}$

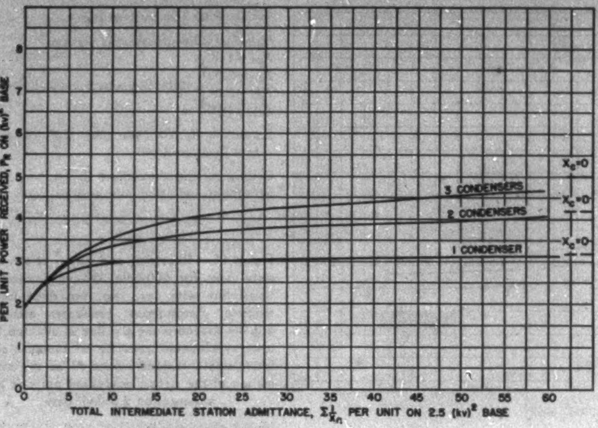


Figure 16. Receiving end power at steady-state limit for 600-mile 60-cycle conventional line with one, two, and three intermediate synchronous condenser stations, equally spaced along the line, as affected by total intermediate station capacity expressed as the summation of the intermediate station equivalent per unit admittance on 2.5 (kv^2) base

$x=0.8$ ohm per mile
$r=0.08$ ohm per mile
$b=5.2\times10^{-6}$ mho per mile
$X_G=40$ per cent [2.5 (kv^2) base]
$X_M=25$ per cent [2.5 (kv^2) base]
1—One intermediate condenser station
Line in two 300-mile sections
$E_1=E_2=E_{c1}=1.0,\ X_c=X_{c1}$
2—Two intermediate condenser stations
Line in three 200-mile sections
$E_1=E_2=E_{c1}=E_{c2}=1.0,\ X_c=X_{c1}=X_{c2}$
3—Three intermediate condenser stations
Line in four 150-mile sections
$E_1=E_2=E_{c1}=E_{c2}=E_{c3}=1.0,\ X_c=X_{c1}=X_{c2}=X_{c3}$

Regulator excitation system amplification factor, $\mu_1=20.0$
Stabilizer amplification factor, $\mu_s=4.0$
Stabilizer time constant, $T_s=176$ radians or 0.466 second

Appendix III. Characteristic System Impedances

Kva base $=5.0\times(\text{kv})^2$
Fault on (3-phase near sending end high-voltage bus)

$Z_{11}=0.015+j0.52=0.52\ \underline{/88.3^\circ}$

$Z_{22}=0.632\ \underline{/84.29^\circ}$

A resistance of 0.015 was assumed for the sending-end terminal impedance during the fault on period to approximate the 3-phase short-circuit loss.

Faulted Line Section Cleared

$Z_{11}=1.39\ \underline{/85.0^\circ}\quad \alpha_{11}=4.2^\circ$

$Z_{22}=1.69\ \underline{/85.0^\circ}\quad \alpha_{22}=5.0^\circ$

$Z_{12}=1.22\ \underline{/86.4^\circ}\quad \alpha_{12}=3.6^\circ$

Appendix IV. System Equations

$$4\pi fH_1\frac{d^2\delta_1}{dt^2}=T_{m1}-T_{e1} \tag{13}$$

$$4\pi fH_2\frac{d^2\delta_2}{dt^2}=T_{m2}-T_{e2} \tag{14}$$

$$T_{e1}=T_{d11}\frac{d\delta_1}{dt}-T_{d12}\frac{d\delta_2}{dt}+T_{s1} \tag{15}$$

$$T_{e2}=-T_{d21}\frac{d\delta_1}{dt}+T_{d22}\frac{d\delta_2}{dt}+T_{s2} \tag{16}$$

$$i_{d1}=\frac{e_{q1}}{Z_{11}}\cos\alpha_{11}-\frac{e_2}{Z_{12}}\cos(\delta_{12}-\alpha_{12}) \tag{17}$$

$$i_{q1}=\frac{e_{q1}}{Z_{11}}\sin\alpha_{11}+\frac{e_2}{Z_{12}}\sin(\delta_{12}-\alpha_{12}) \tag{18}$$

$$T_{s1}=i_{q1}e_{q1} \tag{19}$$

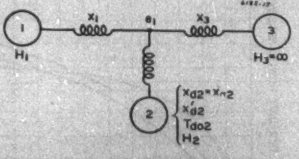


Figure 17. Three-machine system analyzed in Appendix I

Machine 1—Generator
Machine 2—Intermediate synchronous condenser
Machine 3—Infinite inertia receiver system

$$T_{s2}=\frac{e_2^2}{Z_{22}}\sin\alpha_{22}-\frac{e_2e_{q1}}{Z_{12}}\sin(\delta_{12}+\alpha_{12}) \tag{20}$$

$$e_{dt}=-i_{q1}(x_q+x_t) \tag{21}$$

$$e_{qt}=e_{q1}'-i_{d1}(x_d'+x_t) \tag{22}$$

$$e_t=\sqrt{e_{dt}^2+e_{qt}^2} \tag{23}$$

$$\frac{de_{q1}'}{dt}=\frac{1}{T_{do}}(E_{f1}-e_{f1}) \tag{24}$$

$$e_{q1}'=e_{q1}-i_{d1}(x_q-x_d') \tag{25}$$

$$e_{f1}=e_{q1}+i_{d1}(x_d-x_q) \tag{26}$$

$$E_{f1}''=f(E_{f1}') \tag{27}$$

$$\begin{aligned}&=E_{f1}'-E_{f10}'\\&=-\mu_1(e_t-e_{t0})-B\end{aligned} \tag{28}$$

$$B=\frac{\mu_sT_sp}{T_sp+1}E_{f1}$$

$$\frac{dE_{f1}}{dt}=\frac{1}{T_e}(E_{f1}''-E_{f1}) \tag{29}$$

Appendix V. Differential Analyzer Equations (Derived From Appendix IV)

$$\frac{d\delta_1}{dt}=\frac{1}{4\pi fH_1}\Big\{T_{m1}t-T_{d11}\delta_1+T_{d12}\delta_2-\int i_{q1}d\int e_{q1}dt\Big\} \tag{30}$$

$$\frac{d\delta_2}{dt}=\frac{1}{4\pi fH_2}\left\{\left(T_{M2}-\frac{e_2^2}{Z_{22}}\sin\alpha_{22}\right)t+T_{d21}\delta_1-T_{d22}\delta_2+\frac{e_2}{Z_{12}}\int\sin(\delta_{12}+\alpha_{12})d\int e_{q1}dt\right\} \tag{31}$$

$$i_{q1}=\frac{e_{q1}}{Z_{11}}\sin\alpha_{11}+\frac{e_2}{Z_{12}}\sin(\delta_{12}-\alpha_{12}) \tag{32}$$

$$e_{q1}=e_{q1}'+(x_q-x_d')i_{d1} \tag{33}$$

$$e_{q1}'=\frac{1}{T_{do}}\int(E_{f1}-e_{f1})dt \tag{34}$$

$$e_{f1}=e_{q1}+(x_d-x_q)i_{d1} \tag{35}$$

$$i_{d1}=\frac{e_{q1}}{Z_{11}}\cos\alpha_{11}-\frac{e_2}{Z_{12}}\cos(\delta_{12}-\alpha_{12}) \tag{36}$$

$$E_{f1}'-E_{f10}'=-\mu_1(e_t-e_{t0})-B$$

$$B=\mu_sE_{f10}-\frac{1}{T_s}\int B\,dt \tag{37}$$

$$E_{f1}''=f(E_{f1}') \tag{38}$$

$$E_{f1}=\frac{1}{T_e}\int(E_{f1}''-E_{f1})dt \tag{39}$$

$$e_t=\sqrt{e_{dt}^2+e_{qt}^2} \tag{40}$$

$$e_{dt}=-(x_q+x_t)i_{q1} \tag{41}$$

$$e_{qt}=e_{q1}'-(x_d'+x_t)i_{d1} \tag{42}$$

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Functional Analysis of Measurements

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FELLOW AIEE

HISTORY tells us that measurement has been one of the principal foundations upon which the material progress of mankind has been based. It is only by carefully comparing or measuring the unknown in terms of some known quantity or unit that real knowledge is advanced. As our modern way of living increases in complexity and as our knowledge of science and technology expands, measurements and measuring equipment have progressively increased in diversity and in complication. Better measurements have vitally aided our progress in understanding physical laws, in designing and operating new and improved machines, and in automatic control of processes and operations. In all fields of applied arts and sciences, our rich store of knowledge is largely the result of our ability to measure.

Measurements were once simple and relatively few. With the development of practical applications of all branches of science, however, the number of quantities requiring accurate measurement has increased greatly and is now almost without limit. Measurements on a large scale have graduated from the laboratories to our workshops, factories, and even our homes. They have become in one way or another a real and necessary part of our daily lives. Much measuring apparatus has become exceedingly complex. This is caused by a number of factors, including the necessity of performing new measurements inherently more difficult and the need of greater accuracy and sensitivity.

Electrical means for assisting in the measurement of a large number of physical quantities are now quite generally utilized. Their use provides a number of outstanding advantages, such as ready availability of measurement energy and superior flexibility. Moreover, electricity readily lends itself to remote operation of indicators, recorders, or controllers. This trend has further complicated measuring equipment, however, by adding many complex measurement systems and devices that have been designed for a great variety of purposes.

The method of analysis and classification described in this paper concerns itself largely with the measurement of electrical or other physical quantities by methods employing electrical means, but the principles involved are not limited to electrically operated equipment. It was developed after several years of study in order to provide a clear presentation of the various functional parts of any measurement system or device, and to make available a common language of measurement terms and nomenclature for the group of measurement engineers with which the author is associated. Other benefits have accrued from the study which will be outlined later. It is hoped that the material herein presented will be of assistance to all designers and users who are interested in this very important and rapidly expanding field.

General Functional Classification

All measurement systems, simple or complex, can be considered as being composed of not more than three principal functional parts or groupings, namely, a primary detector, one or more intermediate means, and an end device. This conception is illustrated by the diagram shown in Figure 1. The sequence of the measurement is from left to right and all functional steps in the over-all operation logically fall into one of these groups. In many types of measurements, no intermediate means is necessary in order to accomplish the desired result. Very frequently, however, all three general functional classifications are required to analyze a complete system or measurement operation.

END DEVICE AND PRIMARY DETECTOR

A particular quantity is measured to obtain either an indication of its present value, a record showing time of occurrence, or to initiate a control operation. For practical purposes, therefore, a mechanism or assembly that embodies one or all of these functions is called the end device. This represents the result of the measurement and is of primary interest to the user of the equipment. Everything else involved is for the purpose of suitably actuating the end device in proper relation to the quantity being measured.

For classifying purposes, definite and recognized names can be applied to the various types of indicating and recording end devices, and these are well understood. Some further explanation, however, should be made in regard to the control initiator. In those cases where a measurement is made in whole or in part for the purpose of controlling the measured quantity or variable, there is some definite point in the measurement sequence where the control operation can be said to be initiated. This may be in the form of suitable make-and-break contacts, or it

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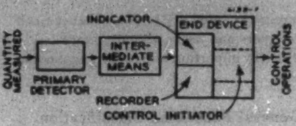


Figure 1. Schematic diagram of general functional classification

may consist of pickup coils and other types of basic elements. The control initiator usually is included in the end device, but it may be associated with the primary detector or intermediate means.

The first operation in the measurement sequence is of particular interest. The basic element or elements here involved must respond quantitatively to the measured quantity or variable in such a manner that the operational result can be used to actuate the end device directly or indirectly. This element or group of elements has been termed the primary detector associated with the measurement apparatus under consideration. It should be emphasized that this conception of a primary detector is purely functional and that some specific basic elements may or may not be primary detectors, depending upon where they are used in the measurement sequence.

As previously indicated, a primary detector and end device may constitute the entire measurement. This is particularly true of the simpler measurement systems. In Table I a number of measurement devices and systems are listed that can be classified suitably in these two functions. In some cases the primary detector may be a self-contained device functioning independently of the actual end device, as illustrated by the first three items in Table I. However, it may constitute an important functioning part of the end device itself, as in the last four devices listed in Table I.

INTERMEDIATE AND AUXILIARY MEANS

In many instances the output or operational result of a primary detector is not suitable to operate the necessary end device either because of its nature or because of its remote location. It, therefore, becomes necessary to convert this output or operational result into some different quantity, to supply measurement energy, or to modify or amplify it in such a manner that it becomes suitable to accomplish the purpose in hand. All such intermediate operations are classified as intermediate means. They are provided for, where necessary, by a proper selection of specific functional basic elements. The location and physical housing of the intermediate means are determined by convenience. It may be included with the primary detector or the end device, or it may consist of separate devices interconnected mechanically or electrically. The thoughtful designer will use a minimum of such transformations or additional operations, since, in general, some measurement precision may be lost in the process. Certainly there is added complexity and cost. Much greater flexibility and sensitivity often can be realized by their use, however, and in many cases the nature of the measurement is such that one or more intermediate means are necessary for its accomplishment. In Table II a number of measurements systems are listed that require intermediate means.

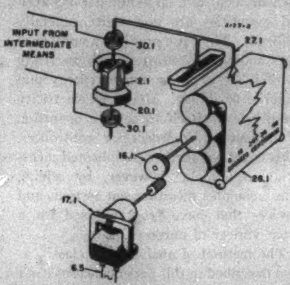


Figure 2. Specific functional classification, D'Arsonval recorder

Basic elements:

- 2.1—Moving coil
- 6.5—A-c standardized frequency source
- 16.1—Gearing
- 17.1—Timing motor
- 20.1—Stationary field magnet
- 26.1—Chart for marking
- 27.1—Curve-drawing means
- 30.1—Restoring-torque springs

Occasionally it may be necessary to change the magnitude of the measured quantity so that its level is better suited to the primary detector. Equipment used for this purpose includes instrument transformers, preamplifiers, and shunts, and is classified in a subordinate grouping as auxiliary means. It is located in front of the primary detector and may be either a separate unit or built in as part of a measurement device. Typical examples are the current and potential transformers used in connection with watthour meters, and also the built-in type of transformers used for conventional thermal watt demand meters.

The auxiliary means constitutes important and necessary equipment for some measurements, but its function is to prepare suitably for the proper operation of the primary detector. It actually does not perform the initial step or quantitative response in the measurement sequence, since it merely changes the level of the measured quantity.

COMMENTS ON GENERAL FUNCTIONAL ANALYSIS

As has been stated, all measuring devices or systems can be classified in accordance with the 3-part functional analysis that has been explained. This has been done for hundreds of different types of measurements and the results tabulated. In general, a rather limited number of functional types of indicating and recording end devices is used. There is a larger list of primary detectors that are available and, undoubtedly, more will be invented. A complete tabulation of the measuring systems that have been analyzed is not included in this paper, but Tables I and II show a few typical cases and will serve to illustrate the manner in which this method can be applied.

The type of analysis shown in Tables I and II provides a means for considering any line or field of measurement apparatus in an orderly manner. Lists of primary detectors, intermediate means, and end devices can be prepared and should be very useful to designers and users of this class of equipment. Such a classification reduces to a relatively few fundamental types what otherwise might appear to be

Table I. General Functional Classification of Simple Measurement Devices and Systems

Measurement Device or System	Quantity Measured	Primary Detector	End Device
Temperature meter (thermocouple type)	Temperature	Thermocouple	D'Arsonval indicator
Light meter	Illumination	Light cell	D'Arsonval indicator
Aircraft synchroscope	Aircraft-engine synchronism	A-c generator	Rotating field and vane indicator
Directional gyroscope	Direction	Horizontal axis gyro	Directional gyro indicator
Interval timer	Time	Timing motor	Time switch
D-c ammeter	Direct current	D'Arsonval mechanism	D'Arsonval indicator
A-c wattmeter	A-c power	Electrodynamic mechanism	Electrodynamic indicator

an almost endless variety of measurement apparatus.

Table II. General Functional Classification of Systems Requiring Intermediate Means

Measurement Systems	Quantity Measured	Primary Detector	Intermediate Means	End Device
Temperature meter (resistance type)	Temperature	Resistance temperature detector	Resistance bridge	Moving-magnet ratio indicator
Photoelectric temperature recorder	Temperature	Thermocouple	Galvanometer, optical transmission, electronic amplifier	D'Arsonval recorder
Remote indicating compass	Direction	Compass mechanism	Induced second harmonic transmission	Induced second harmonic indicator
Liquid-level gauge (capacitance type)	Liquid quantity	Capacitor	Electronic oscillator	Moving-magnet ratio indicator
Watt-demand recording system	A-c power demand	Watthour-meter mechanism	Electric-impulse transmission	Demand recorder

Specific Functional Classification

The term basic element has been used in the text as indicating that measurement devices are composed of certain units which perform specific and definite functions in the measurement sequence. It is quite evident that the classification into general functions does not supply sufficient information to obtain a clear and precise understanding as to the real nature of the various devices under consideration. By further analyzing these devices, therefore, and identifying each specific functional operation in the measurement chain, some very interesting results are obtained. For example, a typical D'Arsonval recorder may be broken down into components or units which perform necessary and distinct steps in the sequence of the measurement operation, as shown in the schematic diagram, Figure 2. There are actually eight functional units as listed that are essential to complete the over-all operation of recording the value of the quantity measured on the chart. For the purposes of exact nomenclature, the units that perform such specific and distinct functions are termed basic elements, and represent the smallest steps in the measurement sequence that it is convenient to consider. These basic elements in a sense may be termed the "building blocks of measurement."

It is now possible to analyze each device or assembly in any measuring apparatus or system and break it down into the basic elements of which it is composed. This has been done for all of the measuring equipment that has been considered from the standpoint of the general functional analysis previously described, and the basic elements identified were listed and classified. In Table III all basic elements thus far considered are grouped into 34 general types, and these expanded into 92 specific types. A decimal index system is used for convenience of reference and for adding new elements as may be required. In preparing diagrams and tabulations, the number of the specific type is always shown, since this definitely identifies the function involved.

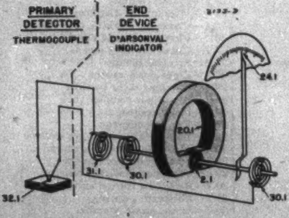


Figure 3. Temperature meter, thermocouple type

Basic elements:

- 2.1—Moving coil
- 20.1—Stationary field magnet
- 24.1—Moving pointer, fixed scale
- 30.1—Restoring-torque springs
- 31.1—Bimetallic actuator
- 32.1—Contact thermocouple

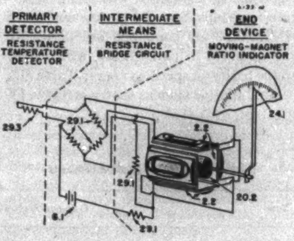

PRIMARY DETECTOR
RESISTANCE TEMPERATURE DETECTOR
INTERMEDIATE MEANS
RESISTANCE BRIDGE CIRCUIT
END DEVICE
MOVING-MAGNET RATIO INDICATOR
29.3
29.1
2.2
24.1
29.1
2.2
20.2
6.1
29.1

Figure 4. Temperature meter, resistance type

Basic elements:

- 2.2—Fixed coils
- 6.1—D-c source
- 20.2—Moving-field magnet
- 24.1—Moving-pointer, fixed scale
- 29.1—Fixed resistors
- 29.3—Temperature sensitive resistor

While only 92 specific types are needed to analyze the measuring equipment thus far considered, it is realized that this list cannot be considered as complete. Additions will have to be made as new types are encountered in existing apparatus, and also as new inventions are made. The number added, however, probably will be small, since experience has shown that nearly all new measuring systems analyzed fit into the pattern of classification that has been established.

It is quite feasible to subdivide further these specific types of basic elements into major and minor variations for purposes of classifying actual physical constructions that are used. The decimal system lends itself to this extension very nicely, and, by this means, types of construction, material used, and actual size and shape may be identified with particular decimal numbers. This actually has been done for some of the basic elements listed and is being extended, since it provides an excellent system for purposes of systematic filing and indexing of data.

Complete Analysis and Classification

It is now possible to show on one schematic diagram a complete measurement system utilizing the methods of analysis that have been described. This provides a standard method of presentation and also a basis for naming, listing, and statistically analyzing all the functional parts of which measurement apparatus is composed, namely, basic elements, primary detectors, intermediate means, and end devices. The secondary classification of auxiliary means can be noted and listed also, as required. In making such an analysis, care should be taken to follow the exact meaning of these terms as closely as possible. Definitions of these and other measurement terms frequently used have been carefully worked out and are given in the following section.

Some of the measurement systems or devices that have been previously mentioned now will be shown in diagrammatic form by combining the general and specific functional classification. Each significant step in the measurement sequence is clearly indicated by identifying the basic elements involved and marking them in accordance with the decimal numbers shown in Table III. The order of the tabulation, in general, corresponds to the flow of measurement energy from the first operation to the end result.

The measurement systems or devices

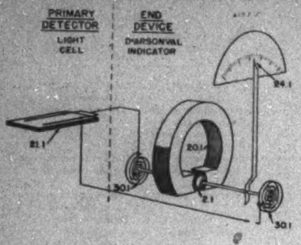


Figure 5. Lightmeter

Basic elements:

- 2.1—Moving coil
- 20.1—Stationary field magnet
- 21.1—Photovoltaic cell
- 24.1—Moving-pointer fixed scale
- 30.1—Restoring-torque spring

selected for illustrating the complete functional analysis method are shown in Figures 3 through 10, and their principles of operation should be self-explanatory if the foregoing material has been carefully considered.

The complete analysis and classification data can be tabulated in compact form for purposes of reference and statistical study. This has been done for a large part of the equipment involved in the electrical measurement field, including apparatus for measuring both electrical and nonelectrical quantities. It is not feasible to show the complete results of this study in this paper, but a skeleton table has been prepared, Table IV. This table shows a number of typical measurements and illustrates how such a tabulation can be built up. A column could be added showing control initiator, if desired, and this is done in considering measurements closely connected with control apparatus. In Table IV the general functional groupings are listed by recognized names associated with their respective functions, and the basic elements of which they are composed are shown by the numbers in parentheses, which refer to the listing of specific types in Table III.

A tabulation such as shown in Table IV can be made quite complete for the measurement of electrical quantities which are relatively few in number, but when extended into the field of measurement of nonelectrical quantities by electrical means, the list is very much larger and is constantly increasing. This clearly shows the growing application of electrical methods to the measurement of physical quantities other than electricity.

Separate lists have been made of basic elements, primary detectors, intermediate means, and end devices. Such data are of considerable importance statistically and as a means of ready reference, as will be explained later.

One mental picture given by a tabulation such as shown in Table IV is the very real significance of the term "building blocks of measurement" as applied to basic elements. It is easy to see at a glance those devices that are complicated and have many steps in the measurement sequence, as compared to those that are simple and have few steps. The point also is brought out that the same basic elements are used in various com-

Table III. List of Basic Elements

General Group	Specific Type	General Group	Specific Type
1—Capacitor	1.1—Fixed capacitor 1.2—Varying dielectric capacitor 1.3—Variable capacitor	19—Pendulum	19.1—Free pendulum
2—Coil, air core	2.1—Moving coil 2.2—Fixed coil	20—Permanent magnet	20.1—Stationary field magnet 20.2—Moving field magnet 20.3—Braking, drag or damping magnet 20.4—Motion transmitting magnet 20.5—Holding, lifting, or pulling magnet
3—Coil, ferromagnetic core (See also inductor-15)	3.1—Electromagnet 3 2—Electromagnet stator 3.3—Electromagnet rotor	21—Photo-sensitive cells	21.1—Photovoltaic cell 21.2—Photoconductive cell 21.3—Photoemissive cell
4—Contacts	4.1—Make-and-break contacts 4.2—Commutator	22—Photo-radiation source	22.1—Electric lamp
5—Eddy-current disk sector or member	5.1—Braking disk 5.2—Drag disk	23—Piezoelectric crystal	23.1—Frequency-standard crystal 23.2—Pressure-conversion crystal
6—Electric source	6.1—D-c source (nonstandardized) 6.2—D-c standardized-voltage source 6.3—A-c source (nonstandardized) 6.4—A-c standardized-voltage source 6.5—A-c standardized-frequency source	24—Pointer and scale	24.1—Moving-pointer, fixed-scale 24.2—Fixed-pointer, moving-scale 24.3—Light-beam, fixed-scale
7—Electrodes	7.1—Nonmetallic electrodes 7.2—Metallic electrodes 7.3—Thermoelectric electrodes	25—Pressure converter	25.1—Expanding vessel 25.2—Liquid column 25.3—Microphone diaphragm
8—Electronic tube (See also photosensitive cells-21 and rectifier-28)	8.1—Grid-controlled electronic tube 8.2—Gaseous discharge tube 8.3—Cathode-ray tube	26—Record charts	26.1—Chart for marking or printing 26.2—Photo-sensitized chart
9—Ferromagnetic moving member	9.1—Ferromagnetic vane 9.2—Ferromagnetic plunger or lever	27—Record-making means	27.1—Curve-drawing means 27.2—Printing means 27.3—Light-beam means
10—Float	10.1—Liquid-level float	28—Rectifier	28.1—Blocking-layer rectifier 28.2—Electronic rectifier tube
11—Fluid-quantity converter	11.1—Eccentric-piston pump	29—Resistor	29.1—Fixed resistor 29.2—Variable resistor 29.3—Temperature-sensitive resistor 29.4—Moisture-sensitive resistor 29.5—Strain-sensitive resistor 29.6—Vibration-sensitive resistor
12—Fluid velocity converter	12 1—Orifice 12.2—Pitot tube	30—Spring	30.1—Restoring torque spring 30.2—Positioning spring 30.3—Stress strain member
13—Frequency standard (See also Piezoelectric crystal-23)	13.1—Tuning fork 13.2—Resonant circuit	31—Temperature converter (See also resistor-29 and thermocouple-32)	31.1—Bimetallic actuator 31.2—Vapor tension bulb
14—Gyroscope	14.1—Electrically driven gyro rotor	32—Thermocouple	32.1—Contact thermocouple 32.2—Radiation thermocouple
15—Inductor or reactor	15.1—Fixed inductor 15.2—Variable inductor 15.3—Mutual inductor 15.4—Saturating reactor	33—Transformer	33.1—Instrument transformer 33.2—Coupling transformer 33.3—Saturating core transformer 33.4—Biased transformer 33.5—Power-supply transformer
16—Mechanical linkage	16.1—Gearing 16.2—Lever 16.3—Clutch 16.4—Cam 16.5—Crank 16.6—Pawl and ratchet	34—Weight	34.1—Solid weight
17—Motor	17.1—Timing motor 17.2—Servo motor		
18—Optical component	18.1—Lense 18.2—Mirror 18.3—Filter		

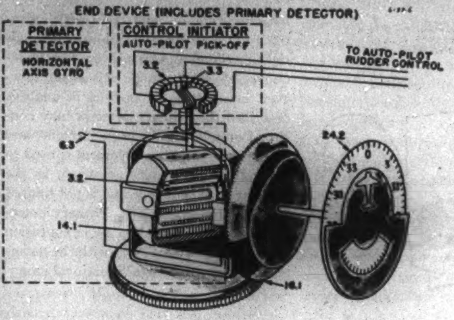

END DEVICE (INCLUDES PRIMARY DETECTOR)
6-29-6
PRIMARY DETECTOR
HORIZONTAL AXIS GYRO
CONTROL INITIATOR
AUTO-PILOT PICK-OFF
TO AUTO-PILOT RUDDER CONTROL
3.2
3.3
6.3
3.2
14.1
16.1
24.2
0
3
33
30

Figure 6. Directional gyroscope

Basic elements:

3.2—Electromagnet stator
3.3—Electromagnet rotor
6.3—A-c source
14.1—Electrically driven gyro rotor
16.1—Gearing
24.2—Fixed-pointer, moving scale

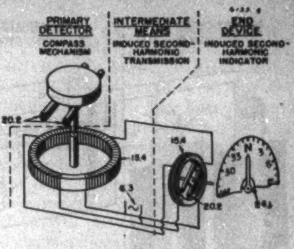

PRIMARY DETECTOR
COMPASS MECHANISM
INTERMEDIATE MEANS
INDUCED SECOND-HARMONIC TRANSMISSION
END DEVICE
INDUCED SECOND-HARMONIC INDICATOR
6-33-8
20.2
15.4
6.3
15.4
20.2
24.1
N
3
33
30
6

Figure 8. Remote indicating compass

Basic elements:

6.3—A-c source
15.4—Saturating reactor
20.2—Moving field magnet
24.1—Moving-pointer, fixed scale

binations to accomplish widely different results.

Definitions of Measurement Terms

It will be helpful at this point to give specific definitions for the various terms that are used to describe the proposed method of analysis. These definitions are carefully worded and are intended to facilitate the actual analysis and classification of measurement devices or systems. Some of these terms are new ones designating somewhat new conceptions, and others are terms that are in general use but to which it is necessary to ascribe more specific meanings.

New terms, which have been selected specifically for the functional analysis and classification of measurements, are as follows:

A primary detector is the first basic element or group of elements that responds quantitatively to the quantity measured and performs the initial measurement operation. A primary detector performs the initial conversion or control of measurement energy and does not include transformers, amplifiers, shunts, resistors, and so forth, when these are used as auxiliary means.

An end device is the assembly of basic elements that responds quantitatively to the quantity measured and performs the final measurement operation. An end device performs the final conversion of measurement energy to an indication, record, or the initiation of control.[*]

The intermediate means includes all basic elements that are used to perform necessary and distinct operations in the measurement sequence between the primary detector and the end device. The intermediate means, where necessary, adapts the operational results of the primary detector to the input requirements of the end device.

A measurement component is a general term applied to parts or subassemblies that are used primarily for the construction of measurement apparatus. It is used to denote those parts made or selected specifically for measurement purposes and does not include standard screws, nuts, insulated wire, or other standard materials.

A basic element is a measurement component or group of components that performs one necessary and distinct function in a sequence of measurement operations. Basic elements are single purpose units and provide the smallest steps into which the measurement sequence can be classified conveniently.

An auxiliary means is a basic element or group of elements which changes the magnitude but not the nature of the quantity being measured to make it more suitable for the primary detector. In a sequence of measurement operations, it usually is placed ahead of the primary detector.

Control initiation is the function introduced into a measurement sequence for the purpose of regulating any subsequent control operations in relation to the quantity measured. The basic element or group of elements comprising the control initiator usually is included in the end device but may be associated with the primary detector or the intermediate means.

Existing terms, which are specifically defined for use in the functional analysis and classification of measurements, are as follows:

A measurement device is an assembly of on

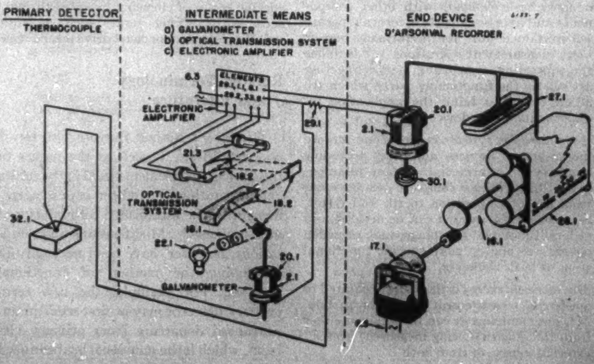
---

[*] It should be noted that end devices, as defined under new terms, are frequently, but not always, complete measurement devices in themselves, since they often are built in with all, or part, of the intermediate means or primary detectors to form separate self-contained units.

Figure 7. Photoelectric temperature recorder

Basic elements:

1.1—Fixed capacitors
2.1—Moving coil
6.3—A-c source
6.5—A-c standardized-frequency source
8.1—Grid-controlled electronic tubes
16.1—Gearing
17.1—Timing motor
18.1—Lenses
18.2—Mirrors
20.1—Stationary field magnet
21.3—Photoemissive cells
22.1—Electric lamp
26.1—Chart for marking
27.1—Curve-drawing means
28.2—Electronic rectifier tube
29.1—Fixed resistors
30.1—Restoring torque spring
32.1—Contact thermocouple
33.5—Power supply transformer


PRIMARY DETECTOR
THERMOCOUPLE
INTERMEDIATE MEANS
a) GALVANOMETER
b) OPTICAL TRANSMISSION SYSTEM
c) ELECTRONIC AMPLIFIER
END DEVICE
D'ARSONVAL RECORDER
6-33-7
6.3
ELEMENTS
29.1, 1.1, 8.1
28.2, 33.5
ELECTRONIC AMPLIFIER
29.1
21.3
18.2
OPTICAL TRANSMISSION SYSTEM
18.2
18.1
22.1
20.1
2.1
GALVANOMETER
32.1
20.1
2.1
30.1
27.1
26.1
16.1
17.1
6.5

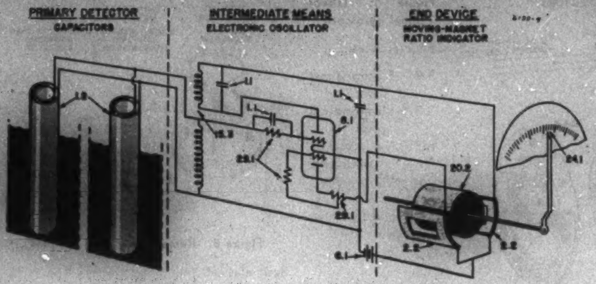


Figure 9. Liquid-level gauge, capacitance type

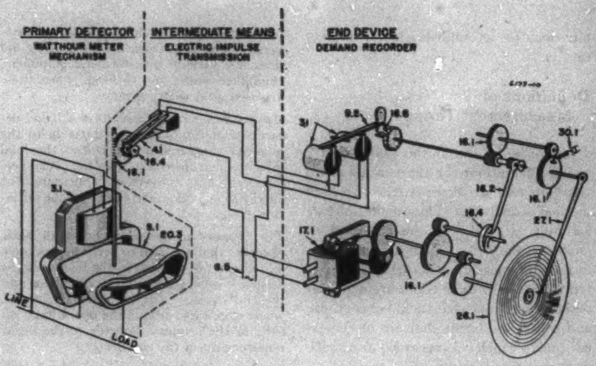
Basic elements:

1.1—Fixed capacitors
1.2—Varying dielectric capacitor
2.2—Fixed coils
6.1—D-c source
8.1—Grid-controlled electronic tube
15.3—Mutual inductor
20.2—Moving field magnet
24.1—Moving-pointer, fixed scale
29.1—Fixed resistors

Figure 10 (right). Watt-demand recording system

Basic elements:

3.1—Electromagnet
4.1—Make-and-break contacts
5.1—Eddy current disk
6.5—A-c standardized-frequency source
9.2—Ferromagnetic lever
16.1—Gearing
16.2—Lever
16.4—Cams
16.6—Pawl and ratchet
17.1—Timing motor
20.3—Braking magnet
26.1—Chart for marking
27.1—Curve-drawing means
30.1—Restoring torque spring



or more basic elements with other components and necessary parts to form a separate self-contained unit for performing one or more measurement operations. It includes the protecting, supporting, and connecting as well as the functioning parts, which are all necessary to fulfill the application requirements of the device.

A measurement system consists of one or more measurement devices and any necessary auxiliary or intermediate means interconnected to perform a complete measurement from the first operation to the end result. A measurement system can be divided into general functional groupings, each of which consists of one or more specific functional steps or basic elements.

Measurement energy is the energy required to operate a measurement device or system. Measurement energy can be obtained either from the quantity being measured, from an external source, or from both.

Measurement equipment is a general term applied to any assemblage of measurement components, devices, apparatus, or systems.

A measurement mechanism is an assembly of basic elements and immediate supporting parts for performing a mechanical operation in the sequence of measurement. For example, it may be a group of components required to effect the proper motion of an indicating or recording means and does not include such parts as bases, covers, scales, and accessories. It also may be applied to a specific group of elements by substituting a suitable qualifying term, such as time switch mechanism or chart drive mechanism.

Related Terminology and Literature

Wherever practical, the foregoing terms and definitions, as well as the names of the devices and elements used in the tables and illustrations, have been kept in agreement with the established terminology of measurement.[1] Modifications and adaptations, however, have been necessary in developing the method of functional analysis. For example, the new term primary detector may at first seem an unwarranted departure from primary element, which is the name that has been used for some time in connection with industrial process instrumentation for the initial functional unit in a measurement sequence.[2,3] This change was made, however, because in the method of analysis described the primary detector may consist of more than one basic element, and because the term basic element is used to cover functional units anywhere in the measurement sequence and is not limited to the initial operation. Primary detector, therefore, is used to avoid confusion and to be more consistent with the names of the other two general functional groupings, the intermediate means and the end device.

Functional analyses of measurement and automatic control systems have appeared previously in technical literature.[2,4,5] While these analyses emphasize the value of considering measurements from the functional point of view, they have been limited, in general, to only a part of the whole measurement field and have been made to facilitate detailed discussions of the particular part involved. The functional analysis method presented herein is intended to be universally applicable to every device and system in the measurement field.

Uses of Method and Conclusions

The method of analysis and classification of measurements that has been described was developed to provide an orderly approach to the problem of understanding and properly classifying the fundamentals of modern measurement systems. The many ramifications and the apparent complexity of this type of equipment make it imperative that terms be standardized more specifically than at present, and a co-ordinated nomenclature, therefore, is suggested.

The principal beneficial results obtained from this study may be summarized as follows.

Clearer Understanding

It seems obvious that a useful medium is made available to depict or illustrate individual measurement systems or devices, showing in a systematic manner and in the proper sequence just how they function. As a corollary, a specific use of words and phrases connected with measurement leads to verbal or written technical treatment that can be understood easily and commonly. It is hoped that this may have some helpful influence on future standardization activities.

By identifying common functional elements or assemblies of elements, a convenient method is provided for exploring and evaluating the whole measurement field and determining fundamental relationships in equipment having widely different applications. It is also of value in studying various designs of measurement apparatus intended for the same purpose and determining fundamental differences and similarities. This all can be accomplished by carefully listing the basic elements, primary detectors, intermediate means, end devices, auxiliary means, or control initiators of which the apparatus under consideration is composed.

Co-ordination and Standardization

As a result of a clearer understanding of the exact nature and purpose of the various functional parts of a measurement system or device, engineers can co-ordinate more effectively the design of specific basic elements, parts, or mechanisms. Co-ordination here is used in the sense of insuring that specific parts having a particular function be used as widely as possible in different apparatus wherever that function is required. Wasteful duplication thus is prevented by minimizing minor variations in parts or mechanisms intended to perform the same or very similar operations.

The interrelation of basic elements, primary detectors, and other functional parts of a system has been emphasized in connection with Table IV. This also is indicated clearly by the fact that only 92 basic elements (Table III) are required to build up many hundreds of greatly diversified systems and devices. Of course

Table IV. Partial Listing of Functional Analysis and Classification of Measurements

Quantity Measured	Primary Detector		Intermediate Means		End Device	
	Name	Basic Elements	Name	Basic Elements	Name	Basic Elements
Group I—Electrical quantities						
D-c amps and volts	D'Arsonval mechanism	(2.1)(20.1)	None	—	D'Arsonval indicator	(2.1)(20.1)(30.1)(24.1)
	D'Arsonval mechanism	(2.1)(20.1)	None	—	D'Arsonval recorder	(2.1)(20.1)(30.1)(17.1)(6.5)(16.1)(26.1)(27.1)
A-c amps and volts	Moving iron mechanism	(3.1)(9.1)	None	—	Moving iron indicator	(3.1)(9.1)(30.1)(24.1)
	Thermocouple converter	(29.3)(32.1)	None	—	D'Arsonval indicator	(2.1)(20.1)(30.1)(24.1)
	Electronic rectifier	(29.1)(1.1)(8.1)(28.2)(33.5)(6.3)	None	—	D'Arsonval indicator	(2.1)(20.1)(30.1)(24.1)
A-c power	Electrodynamic mechanism	(2.1)(2.2)	None	—	Electrodynamic indicator	(2.1)(2.2)(30.1)(24.1)
A-c energy	Watthour meter mechanism	(3.1)(5.1)(20.3)	None	—	Watthour meter	(3.1)(5.1)(20.3)(16.1)(24.1)
A-c power demand	Watthour meter mechanism	(3.1)(5.1)(20.3)	Elec impulse transmission	(16.1)(16.4)(4.1)(6.5)	Demand recorder	(3.1)(9.2)(16.6)(16.1)(17.1)(16.2)(16.4)(30.1)(27.1)(26.1)
	Bimetallic actuator	(33.1)(31.1)	None	—	Thermal demand indicator	(33.1)(31.1)(16.1)(24.1)
Totalized a-c power demand	Watthour meter mechanism	(3.1)(5.1)(20.3)	Elec impulse transmission Totalizing relay	(16.1)(16.4)(4.1)(6.5) (3.1)(9.2)(16.6)(16.1)(16.4)(4.1)(6.5)	Demand recorder	(3.1)(9.2)(16.6)(16.1)(17.1)(16.7)(16.4)(30.1)(27.1)(26.1)
Group II—Nonelectrical quantities						
Speed	A-c generator	(20.2)(3.2)	Synchronous motor	(3.2)(20.2)	Magnetic drag disk indicator	(20.3)(5.2)(30.1)(24.1)
	A-c generator	(9.1)(20.1)(3.2)	Frequency sensitive rectifying network	(33.3)(28.1)(29.1)	D'Arsonval indicator	(2.1)(20.1)(30.1)(24.1)
Aircraft-engine synchronism	A-c generator	(20.2)(3.2)	None	—	Rotating field and vane indicator	(3.2)(9.1)(24.1)
Temperature	Resistance temperature detector	(29.3)	Resistance bridge	(29.1)(6.1)	Moving mag ratio indicator	(29.1)(2.2)(20.2)(24.1)
	Thermocouple	(32.1)	None	—	D'Arsonval indicator	(2.1)(20.1)(31.1)(30.1)(24.1)
	Thermocouple	(32.1)	Galvanometer Optical transmission Electronic ampl	(2.1)(20.1) (22.1)(18.1)(18.2)(21.3) (29.1)(1.1)(8.1)(28.2)(33.5)(6.3)	D'Arsonval recorder	(2.1)(20.1)(30.1)(17.1)(6.5)(16.1)(26.1)(27.1)
Pressure	Bellows mechanism	(25.1)	Induced 2nd harmonic transmission	(20.2)(15.4)(6.3)	2nd harmonic ratio indicator	(15.4)(20.2)(20.1)(24.1)
Flow	Pump mechanism	(11.1)(20.4)	Selsyn transmission	(4.2)(6.1)(29.1)(3.2)(20.2)	Magnetic drag disk indicator	(20.3)(5.2)(30.1)(24.1)(16.1) (24.2)
Liquid level	Float mechanism	(10.1)(16.1)(20.4)	D-c Selsyn transmission	(29.2)(6.1)	D-c Selsyn indicator	(3.2)(20.2)(24.1)
	Capacitor	(1.2)	Elec oscillator	(15.3)(1.1)(29.1)(8.1)(6.1)	Moving mag ratio indicator	(2.2)(20.2)(24.1)
Direction	Horizontal axis gyro	(14.1)(3.2)(6.3)	None	—	Directional gyro indicator	(14.1)(3.2)(6.3)(16.1)(24.2)
	Compass mechanism	(20.2)	Induced 2nd harmonic transmission	(15.4)(6.3)	2nd harmonic indicator	(15.4)(20.2)(24.1)
Time	Timing motor	(17.1)(6.5)	None	—	Time switch	(16.1)(24.2)(16.4)(16.2)(30.2)(4.1)
Light	Light cell	(21.1)	None	—	D'Arsonval indicator	(2.1)(20.1)(30.1)(24.1)

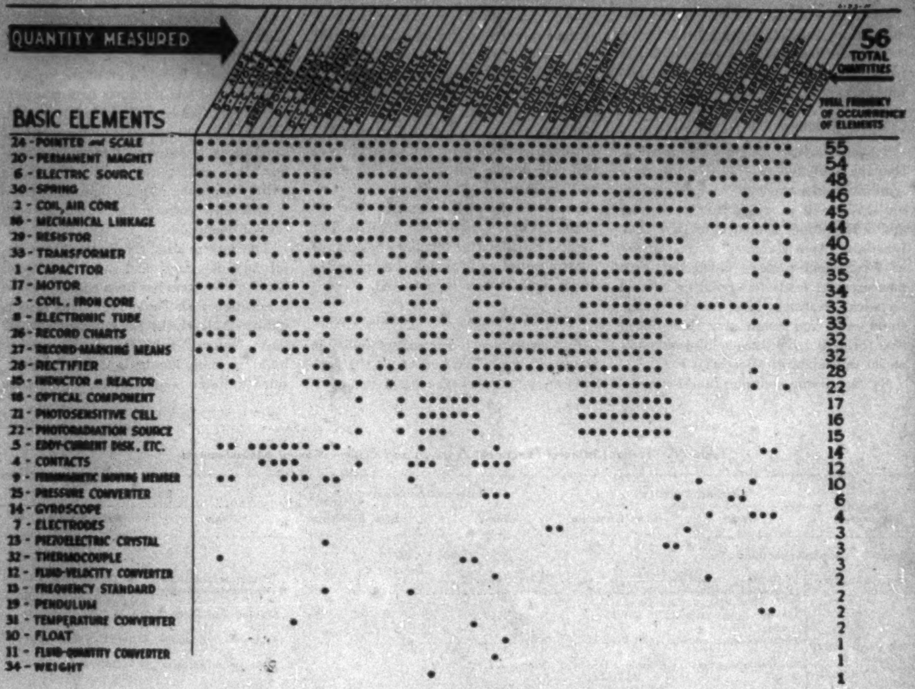


Figure 11. Frequency of occurrence of basic elements

many varieties of these have been developed for specific purposes, probably many more than are economically sound. It would seem that engineers are concerned mostly with designing some modifications of existing basic elements and regrouping them in order to accomplish a desired result. It rarely happens that a really new one is added.

In order to get a rough idea of the extent to which each basic element that has been identified actually is used for various types of measurements, the tabulation shown in Figure 11 has been prepared. Apparatus suitable for measuring 56 of the more important electrical and nonelectrical quantities was analyzed and the necessary basic elements listed. In Figure 11, basic elements involved are shown in the order of their frequency of occurrence. This arrangement is useful in emphasizing those elements that probably require most attention for co-ordination purposes. It is not claimed, however, that this tabulation is complete, nor entirely accurate. It is a generalization and presents a point of view that may be useful.

All progress that can be made in co-ordinating and standardizing functional parts of devices and systems will be, of course, a powerful influence in arriving at better standardization of measuring equipment as a whole.

RESEARCH AND DEVELOPMENT

Well-arranged lists of basic elements and the general functional parts of all known measurement apparatus or equipment are very helpful to research and development engineers. They indicate clearly what is available and where new development or inventions are required. Furthermore, the general viewpoint and over-all understanding that can be gained from an analysis such as outlined enables an engineer to focus his attention directly on a desired specific objective and to see it in proper relation to the general problem in hand and the entire measurement field.

APPLICATION AND SERVICE

In the selection of measurement equipment for a specific purpose, the functional analysis method can be quite helpful in several ways. In the first place, the simplified functional diagrams should aid in understanding and choosing the proper instrumentation and in maintaining it in service. Evaluations and comparisons made possible by the system described aid materially in answering such questions as the following. How direct is the measurement? Is there undue complexity, especially in intermediate means? Are the basic elements inherently simple and reliable? Is the design well co-ordinated? What auxiliary means, or power supply, is necessary? Correct answers to these and other pertinent questions can be expedited if the application engineer is familiar with the idea of functional analysis and uses it carefully as an aid to his investigations.

DATA AND STATISTICS

Considerable use already has been made of the systematic classifications and tabulations of measurement functional parts for purposes of filing and cross-indexing.

Simplicity in Protective Relaying

LLOYD F. HUNT
FELLOW AIEE

SIMPLICITY is of paramount importance when designing and laying out protective relay circuits. These circuits are called upon to operate so infrequently that they should be more reliable than the apparatus or lines they protect. Certain limiting conditions and requirements requested by circuit designers generally cause false tripping and make the circuits less reliable. A simple protective system with all the extra devices left out operates so satisfactorily that the reasons for its good operation are often forgotten. The simple rules concerning protective systems in use on the Southern California Edison System that have made it possible to attain such a high standard service to its customers are reviewed in this paper.

A short study of fault conditions reveals that by far the majority of failures are faults to ground. On very high voltage transmission lines on steel towers ground faults approximate 98 per cent of all faults. On lower voltage transmission lines and higher voltage distribution lines on wood poles, ground faults drop to about 90 per cent. For rotating machines and transformers ground faults reach 98 per cent. Therefore the task for protection engineers is to design protection to isolate ground faults first. If a perfect system of only ground fault protection were used, the service record of relay performances would be an excellent one. It is therefore important that the added protection necessary for faults other than those to ground should be simple.

Conclusions

This paper presents the principles that have guided the application of protective systems over a period of the last 20 years. It is evident from the above-average record of operation that they have paid in equipment saved, and outages and false tripping prevented.

Because all these principles have been based on simplicity and reliability, the results of continued service have justified their practices.

Though many relay applications have been pioneered on the Edison system, the trend has been to apply protection primarily for the majority of faults. Elimination of complicated schemes for protection against less important faults has resulted in a system that has been designed and maintained easily. The record of approximately 99 per cent perfect relay operations over a long period of time indicates the justification of these principles of simplicity and reliability.

Introduction

The Southern California Edison Company, Ltd. has primarily a transmission type system. It is expanding continuously to meet the needs of a very rapidly growing community serving 17,500 square miles in southern California. Its present transmission system is shown in Figure 1. From this diagram it can be seen that the backbone of the system is the 220-kv line starting at Big Creek and Boulder power plants, running through or to the major stations and ending at Long Beach number 3 steam plant.

The load is supplied through the 66-kv subtransmission system. The 66-kv system is divided into many sections. At each major substation where 220 kv is stepped down to 66 kv, the 66-kv system is a unit by itself, not normally paralleled with any other 66-kv group. Even at some of these major substations the 66-kv busses are split to form two of these groups. For example, the 66-kv system at Eagle Rock can be paralleled or it can be separated, each part being supplied from one transformer bank. It is not clearly shown in Figure 1 where the separation of the 66-kv lines are made between 66-kv systems. This is variable according to load conditions. For example, the 66-kv system of Eagle Rock

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Data on primary detectors and end devices, for example, can be indexed in accordance with the quantities to which they respond, as well as by their names on the charts. Intermediate and auxiliary means can be indexed relative to the function they perform and also the specific names assigned to them. Basic elements are specifically and conveniently listed on the chart and each one can be subdivided into detailed variations as far as desired. All technical information available can be so indexed that it is readily identified with the pertinent basic element. These examples will serve to point out the usefulness of this method in organizing and indexing technical information.

General

The benefits realized by the functional analysis and classification of measurement systems and devices that has been described will depend a great deal upon how much it is really studied, understood, and applied in a given organization or industry. In this paper it has been possible to present only a small part of the work that has been done on the subject. It is also likely that the point will be raised that nothing really new or radical has been presented. In one sense this is true. What has been attempted, however, is to collect, analyze, systematize, and organize information, methods and nomenclature that have been used in the art, but used in an unsatisfactory and indefinite manner. In doing this it was found necessary to coin some new words and phrases and to give these, as well as those commonly used, specific meanings and definitions. It is felt, too, that a complete analysis consistently and systematically carried out from the functional viewpoint is new. Actually, it has already proved very useful in a practical manner. Although little thought has been given to the matter, it is also probable that the methods described in this paper can be extended advantageously to apparatus and equipment quite outside of the measurement field.

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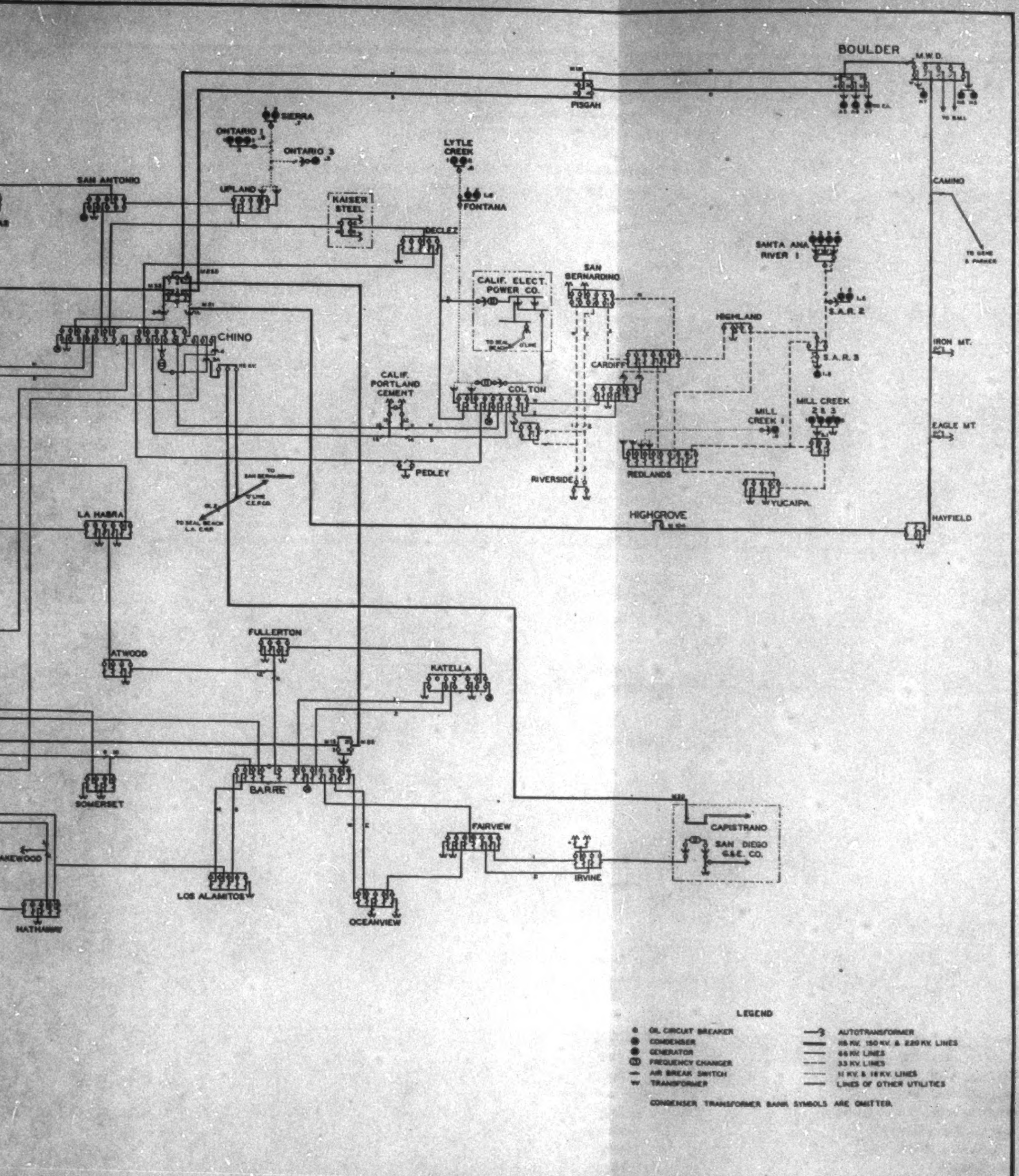
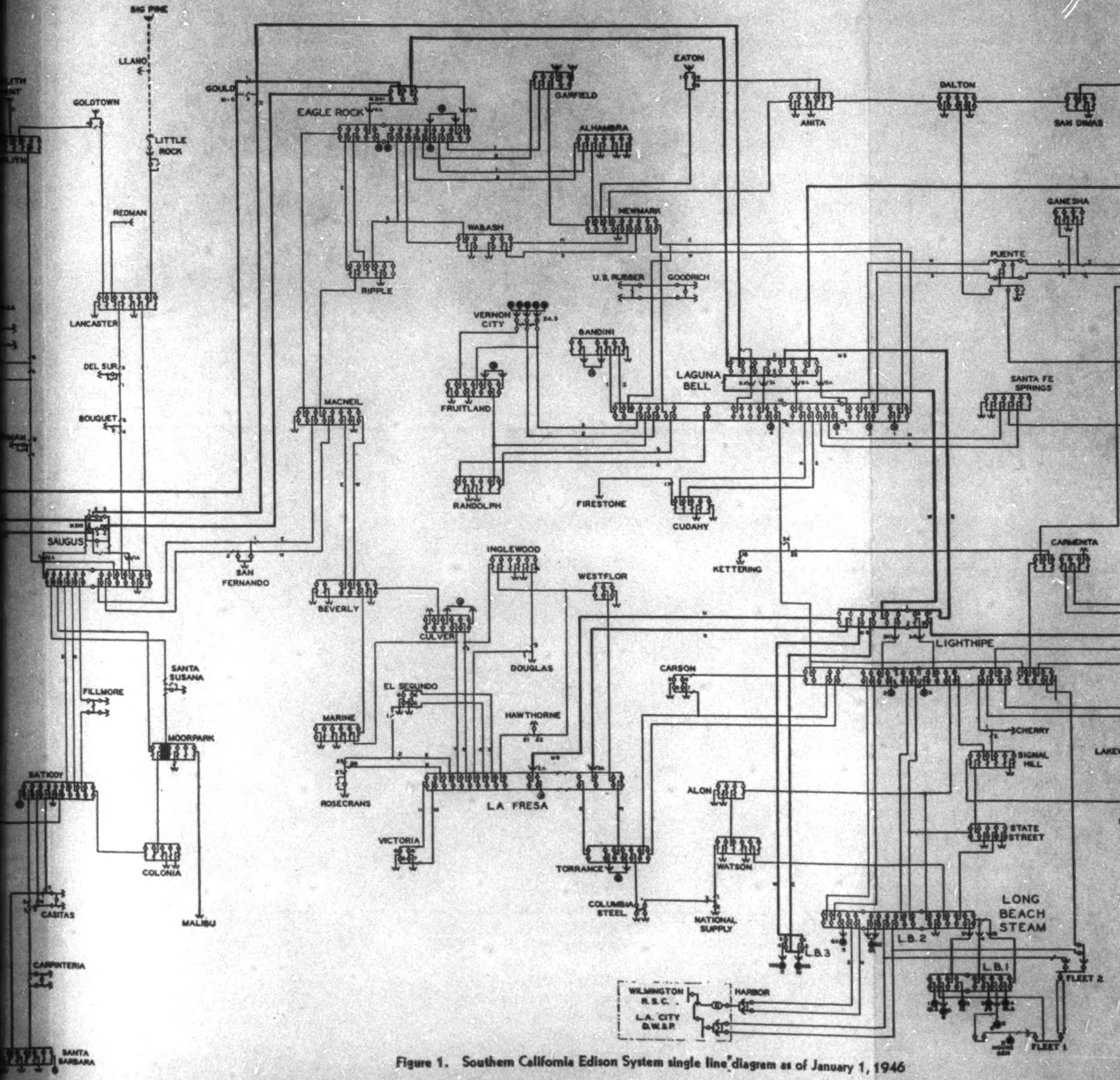

BOULDER
M.W.D.
TO S.M.I.
PISGAH
SIERRA
ONTARIO 1
ONTARIO 3
LYTLE CREEK
FONTANA
SAN ANTONIO
UPLAND
KAISER STEEL
DECLEZ
CAMINO
TO GENE & PARKER
SANTA ANA RIVER 1
CALIF. ELECT. POWER CO.
TO SEAL BEACH
SAN BERNARDINO
S.A.R. 2
HIGHLAND
S.A.R. 3
IRON MT.
CHINO
CARDIFF
CALIF. PORTLAND CEMENT
COLTON
MILL CREEK 1
MILL CREEK 2 & 3
EAGLE MT.
REDLANDS
PEDLEY
RIVERSIDE
YUCAIPA
TO SAN BERNARDINO
C.E.P.CO.
TO SEAL BEACH L.A. CHR
LA HABRA
HIGHGROVE
HAYFIELD
FULLERTON
ATWOOD
KATELLA
SOMERSET
BARRE
CAPISTRANO
SAN DIEGO G.&E. CO.
FAIRVIEW
IRVINE
LAKEWOOD
LOS ALAMITOS
OCEANVIEW
HATHAWAY
LEGEND
OIL CIRCUIT BREAKER
CONDENSER
GENERATOR
FREQUENCY CHANGER
AIR BREAK SWITCH
TRANSFORMER
AUTOTRANSFORMER
115 KV. 150 KV. & 220 KV. LINES
66 KV. LINES
33 KV. LINES
11 KV. & 18 KV. LINES
LINES OF OTHER UTILITIES
CONDENSER TRANSFORMER BANK SYMBOLS ARE OMITTED

SYSTEM AS OF JANUARY 1, 1946


BIG PINE
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GOULD
GOLDTOWN
EAGLE ROCK
GARFIELD
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SAN DIMAS
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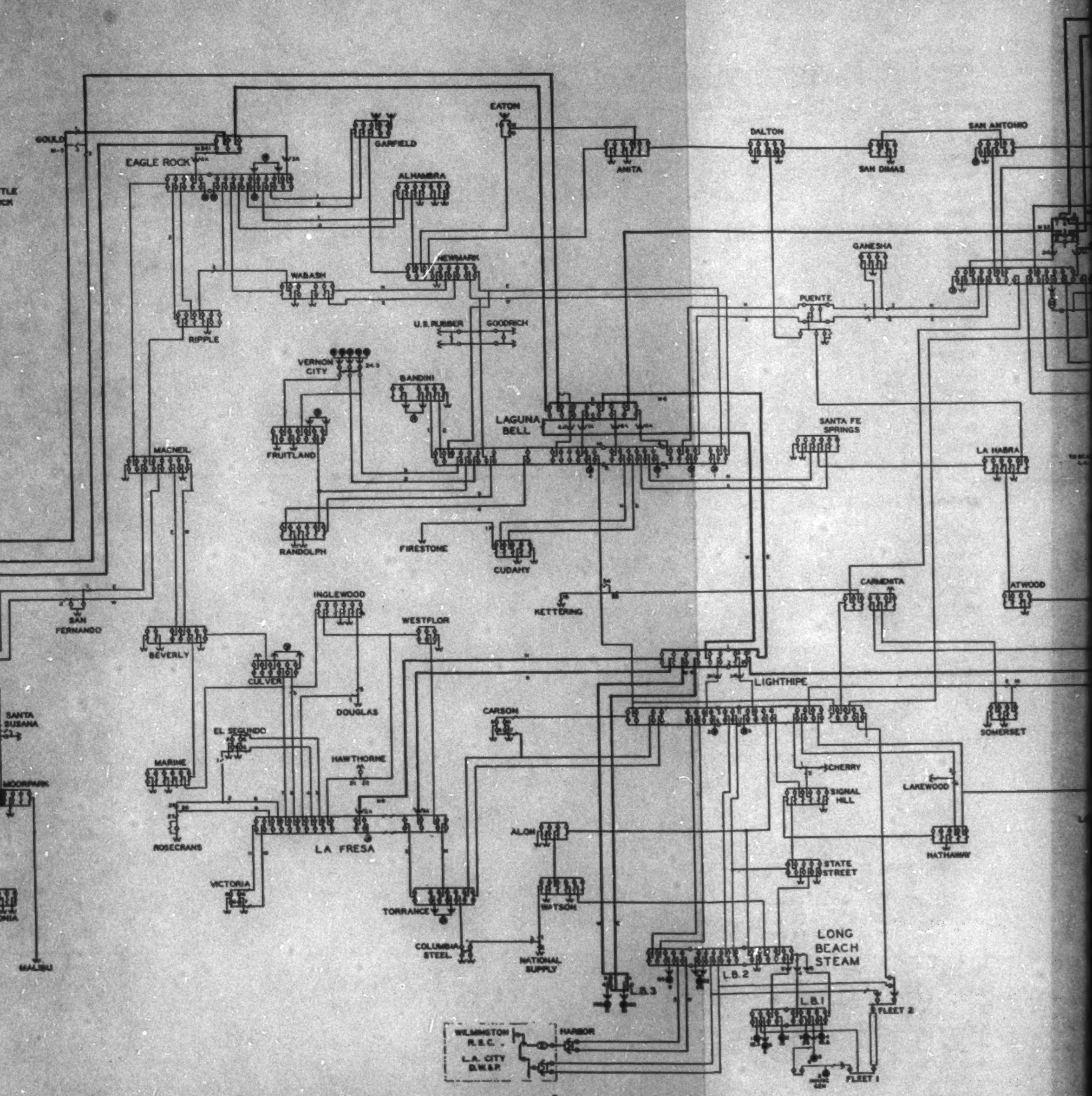
ALHAMBRA
REDMAN
WABASH
NEWMARK
GANESHA
PUENTE
RIPPLE
U.S. RUBBER
GOODRICH
LANCASTER
VERNON CITY
BANDINI
DEL SUR
LAGUNA BELL
SANTA FE SPRINGS
BOUQUET
MACNEIL
FRUITLAND
RANDOLPH
FIRESTONE
CUDAHY
SAUGUS
SAN FERNANDO
INGLEWOOD
KETTERING
CARMENITA
WESTFLOR
BEVERLY
CULVER
LIGHTHIPE
SANTA SUSANA
CARSON
DOUGLAS
FILLMORE
EL SEGUNDO
HAWTHORNE
SCHERRY
MARINE
MOORPARK
SIGNAL HILL
SATICOY
ROSECRANS
ALON
L.A. FRESA
STATE STREET
VICTORIA
COLONIA
TORRANCE
WATSON
CASITAS
COLUMBIA STEEL
NATIONAL SUPPLY
LONG BEACH STEAM
MALIBU
L.B.2
L.B.3
L.B.1
FLEET 2
CARPINTERIA
WILMINGTON R.S.C.
L.A. CITY D.W.&P.
HARBOR
FLEET 1
SANTA BARBARA

Figure 1. Southern California Edison System single line diagram as of January 1, 1946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LIGHTHIPE
DOUGLAS
SOMERSET
SANTA SUSANA
CARSON
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HAWTHORNE
SIGNAL HILL
LAKEWOOD
MARINE
MOORPARK
ROSECRANS
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VICTORIA
TORRANCE
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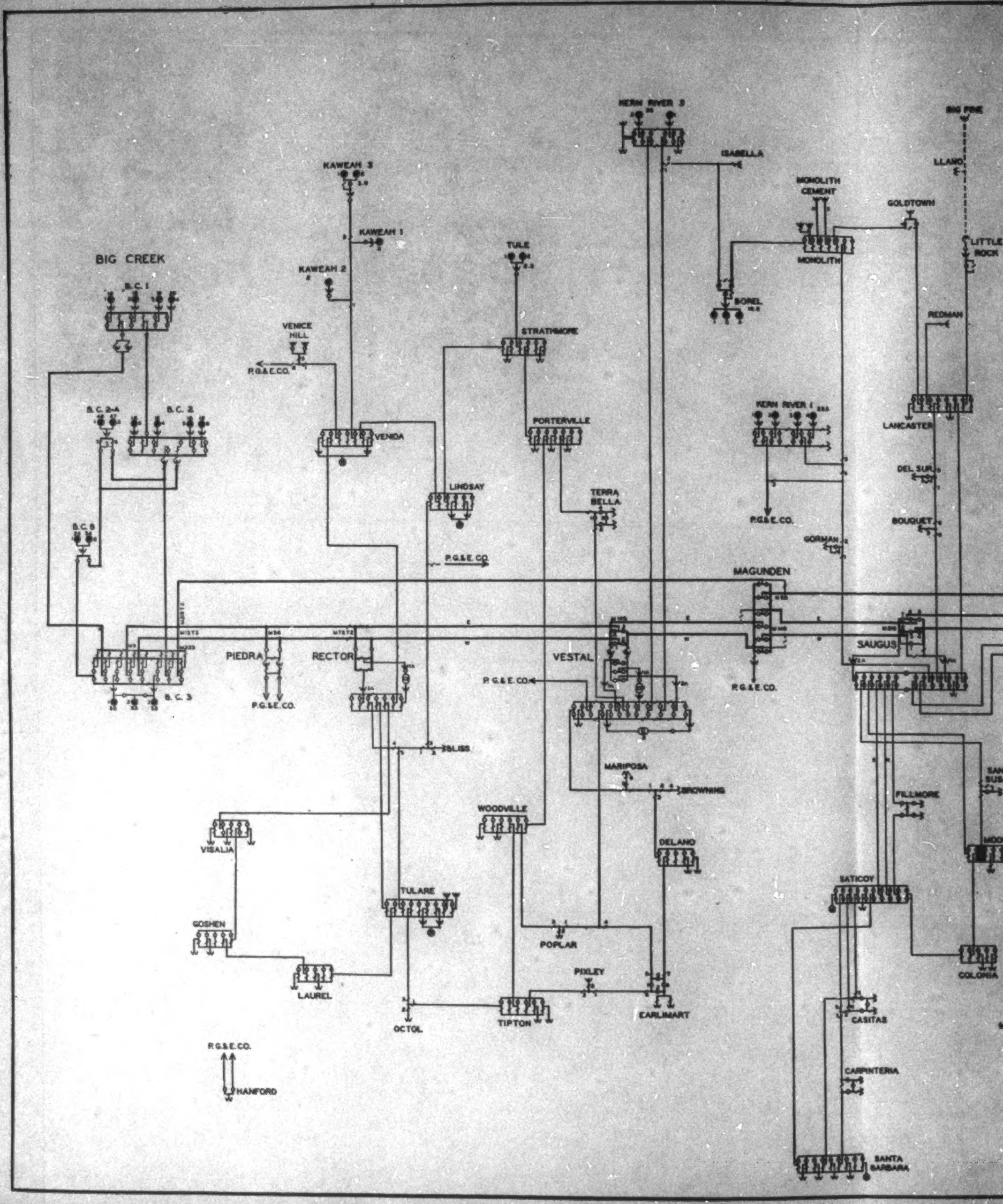
NATIONAL SUPPLY
LONG BEACH STEAM
L.B. 2
L.B. 3
L.B. 1
FLEET 2
MALIBU
WILMINGTON R.S.C.
L.A. CITY D.W.&P.
HARBOR
FLEET 1

Figure 1. Southern California Edison System single line diagram as of January 1, 1946


BIG CREEK
B.C. 1
B.C. 2-A
B.C. 2
B.C. 8
B.C. 3
KAWEAH 3
KAWEAH 1
KAWEAH 2
VENICE HILL
P.G.&E.CO.
VENIDA
TULE
STRATHMORE
PORTERVILLE
LINDSAY
TERRA BELLA
P.G.&E.CO.
PIEDRA
P.G.&E.CO.
RECTOR
VESTAL
P.G.&E.CO.
BLISS
MARIPOSA
BROWNING
WOODVILLE
DELANO
VISALIA
TULARE
GOSHEN
POPLAR
LAUREL
PIXLEY
OCTOL
TIPTON
EARLIMART
P.G.&E.CO.
HANFORD
KERN RIVER 3
ISABELLA
MONOLITH CEMENT
MONOLITH
GOLDTOWN
BOREL
BIG PINE
LLANO
LITTLE ROCK
REDMAN
KERN RIVER 1
P.G.&E.CO.
LANCASTER
DEL SUR
BOUQUET
GORMAN
MAGUNDEN
P.G.&E.CO.
SAUGUS
FILLMORE
SATICOY
COLONIA
CASITAS
CARPINTERIA
SANTA BARBARA

and Laguna Bell normally do not operate in parallel.

Figure 1 shows 1,506 circuit miles of 220-kv lines and 2,130 circuit miles of 66-kv lines supplied by 872,200 kw of hydroelectric generating capacity and 436,250 kw of steam and other generating capacity. The total frequency-changer capacity for 50 to 60 cycles and vice versa is 185,000 kw. Synchronous condenser capacity is 573,000 kva. The maximum peak load to date is 1,025,200 kw. The daily load factor during last year reached as high as 87 per cent, and the yearly load factor was over 70 per cent.

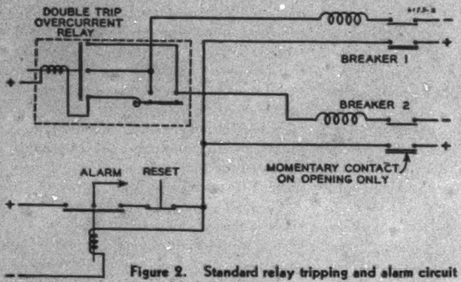


Figure 2. Standard relay tripping and alarm circuit

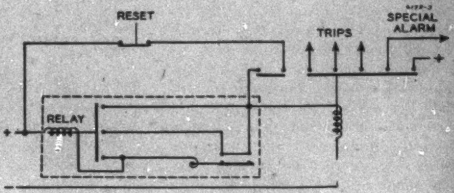


Figure 3. Standard auxiliary relay tripping circuit

Protection Features

It is the duty of large electric systems to supply continuous service to their customers. If a perfect system could be built economically, no relaying systems would be required. This being impossible, it is important that the protection group study the problem from several viewpoints; first, of preventing faults on the system; second, of studying behavior of faults with emphasis placed on troubles that affect the system as a whole; third of directing design of protective systems by which the type and location of faults can be determined and possibly be remedied without de-energizing circuits in which the dropping of load or generation is involved; and fourth, of relaying the smallest portion of the system in order to isolate the fault.

The protection group in the Edison organization has a great deal of influence on system planning, station layout, and design. It has pointed out factors in design which were potential weaknesses and which consequently were avoided. In the layout of stations it is possible to make slight modifications to limit the amount of equipment tripped out in the case of a fault, also to arrange oil circuit breakers to give flexible operation and over-all protection. Recording devices such as automatic oscillographs, Hall recorders and disturbance recorders have been installed at strategic locations.[1] The records of these devices are used for studying methods of better protection. The recorders also are used in the determination of the location of faults. This has aided materially the field forces in checking the condition and damage of the transmission lines.

Protection of Transmission System

Because the prime purpose of protection is to maintain service to customers and to do as little relaying as possible to accomplish this without leaving the facilities in a hazardous condition, it behooves the protection men to use systems of simple circuits.

In the initial application of relays to the Edison system, many individual design engineers took it upon themselves to make up circuits for installations upon which they were working. It was soon found that every job had its distinctive style of connections without any co-ordination of relays. The protection group no longer could be satisfied with studying the behavior of relays but had the duty of dictating types and styles to fit in the system as a whole.

In the original circuit designs the relays operated annunciators, bells, and flags without much regard to tripping circuit breakers. This soon led to a simple standard rule that has continued through the years. That is, the trip circuits of relays are to be the simplest for direct tripping and that only the necessary knife switches be included to facilitate relay testing. Because switches can open from causes other than by means of the trip coil, it was decided, and still is an Edison standard, to have an auxiliary switch on circuit breakers to operate the alarm circuits. Earthquakes and passing trains have caused circuit breakers to trip without signals to the operator under normal setup; however, with the momentary closing of an auxiliary switch upon the opening of the circuit breaker, an alarm is set off. All tripping circuits should be locked in until the auxiliary switch on the circuit breaker has released the trip-coil current. In many locations, as shown on Figure 1, it is necessary to trip two circuit breakers. This is done with double-trip relays. Figure 2 shows schematically the standard tripping and alarm circuits. When more than two circuit breakers are tripped, as in the case of differential protection, an auxiliary relay is used with a sealing-in circuit as shown in Figure 3. This type has a separate reset circuit which is common for tripping all auxiliary relays in the station, thus requiring the operator to use only one reset button for his entire station. A special alarm is used on each auxiliary relay, and thus the operator knows he must reset it before any switching is done. This seems apparent, but the less operators have to remember about such things during system disturbances, the better is the operating record. Many years ago, a hand-reset tripping-relay caused a very long system outage. The operator had no signal to inform him that this relay had not been reset and in the excitement of the trouble he forgot to reset the relay. When the circuit breaker was closed it tripped out immediately, and it was assumed the line (the only 220-kv line left) had a permanent fault. Since this disastrous outage occurred the rule is that all tripping circuits are to be released automatically as in Figure 2, or released manually with an alarm on the relay as in Figure 3.

The relay system on the 220-kv lines is differential from one end to the other, without adjacent section backup, and without overload protection. The word differential is used in the broad sense rather than literally. All of the relays are not connected differentially, but are arranged so as not to trip when faults occur in adjacent sections. The primary protection for all 220-kv line sections, is of instantaneous overcurrent type. All 220-kv terminals are equipped with three phase, and one ground element relay as shown by Figure 4. The ground elements are set to operate on slightly above maximum through ground current. In these cases the ground elements will trip on sequence when the fault is close to one end, and simultaneously when the fault is in the center section. The phase elements are set to operate on just above out-of-step currents. On long sections this setting covers about 40 per cent of the distance from the station. In the shorter sections when all lines are in service these settings generally cover 100 per cent of the line in sequential tripping. To secure tripping for the remaining portion of the lines, or when some lines are out of service, a carrier system of relaying is used when necessary. Otherwise, on long lines where blocking is not essential, the relay scheme of carrier is used without the carrier blocking

The Edison system's first 220-kv carrier-current relay system was started in 1926. This system was very complicated and the record of its operation was unsatisfactory. There were more false tripouts than correct tripouts. In fact, more cases of internal faults were cleared on this line section by the alternate group of relays because it was necessary to have the carrier relay group out for repairs.

The next line section that was equipped with this type of carrier system had a 100 per cent performance record on both internal and external faults, but the record of the very complicated system changed to about 50 per cent operation because of many unnecessary tripouts. This complication, excessive maintenance, time out of service, and replacement costs, required an investigation of a simple and more reliable system. The system which was developed[2] utilizes a lockout carrier or pilot wire and is very simple, because the short circuits on the 220-kv system were mostly ground faults, a few phase-to-phase or phase-to-phase-to-ground, it was believed that simplicity justified leaving out the instantaneous 3-phase fault protection. The phase protection consists of a balanced-current relay for tripping, that is, *A* phase is balanced against *B* phase, *B* phase against *C* phase, and *C* phase against *A* phase. If an unbalance of ten per cent occurs at short-circuit values of current, the relay will trip. It can be seen readily that no complication is needed for guarding against tripping on out-of-step conditions.

For parallel lines, only balanced-current ground relays are used. With high-speed circuit breakers, high-speed balance relays are used and slower relays still are used with slower circuit breakers. This protection has discriminated successfully on all short circuits. With many staged short-circuit tests at Laguna Bell and Saugus substations[3] tied into the system, the whole 220-kv relay setup had an excellent test and no false tripping occurred.

The most difficult work connected with any carrier system is the maintenance of the radio sets. It became necessary to devise a transmitting and receiving set with a minimum of parts, especially capacitors and radio-frequency choke coils. The result was a set employing push-pull oscillator and push-pull power-amplifier stages, using a total of four tubes plus rectifier tubes. These sets proved economical in first cost and operated with reliability.

The terminal protection on the 220-kv system consists of an instantaneous overcurrent relay, carrier, and an auxiliary group consisting of instantaneous overcurrent and time-delay relays. Because there are no normal backup relays, it is all the more important that relays be maintained in good operating order at all times. The auxiliary group of relays is used while the main group of relays is being tested and adjusted. This is necessary because distances between installations are great, and it is desirable when highly trained test engineers are making routine tests that the relays be available for inspection.

The switching from one group of relays to another is required for all terminals when the lines cannot be taken out of service for tests. As mentioned before, in major stations on which the relays trip two breakers, push buttons are supplied so that while testing one switch at a time it may be tripped from the relays being tested.

The 66-kv system is protected by the use of inverse time setting of overload and directional relays. However, at the major supply stations such as Eagle Rock, Laguna Bell, and Lighthipe the 66-kv lines also are equipped with the same 4-element instantaneous-overcurrent relays shown in Figure 4. In these installations they are set for approximately 90 per cent of the line length. With this setting these relays have operated in many cases on the whole line after the switch had opened at the remote end.

Even the simple overcurrent directional relay has three features that make it difficult to apply.

1. It was discovered that much false tripping occurred by changing direction of flow after one end of the faulted line relayed. Study of the problem revealed that it was necessary to govern the overload element by only allowing it to close contact when the direction was correct. In those early days the Edison system built a few sets of directional-controlled relays, which were placed in service at a location where considerable

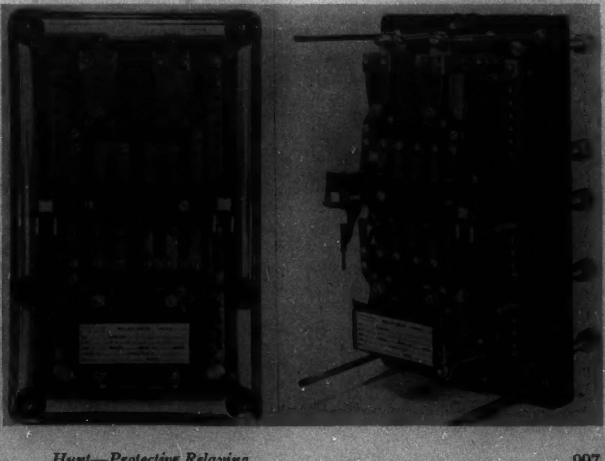


Figure 4. Four-element instantaneous overcurrent relay

difficulty with false tripping had occurred. The results were gratifying. However, it took personal contact with the factory to prove that it could be done and was necessary. It is self sufficient to say that all inverse-directional overcurrent relays are now directionally controlled.

2. The application of single-phase directional overcurrent relays must be watched. If the currents in the unfaulted phases of a ground fault reach values above minimum trip setting of the relays, one of the unfaulted-phase relays on unfaulted lines will be in the direction of trip and will cause unnecessary tripouts.

3. In all directional schemes operated by potential it is possible to have a short circuit that will reduce the voltage to a value so low that the relays will not function. Rigid inspection is required to keep relays in condition to operate at very low voltages. In fact, a very thorough study was made several years ago about changing the 66-kv type of protection. This study revealed that if the directional elements of the relays were modernized according to our requirements, we would have as good an operating system with the existing circuit breakers as possible. As a result the directional elements were modernized with about $35,000 worth of parts. The results have proved the study was right because it is now very seldom that a short circuit occurs which reduces the voltage sufficiently to cause false tripping.

At present a program of rebuilding all 220-kv circuit breakers so that they will range from three to five cycles in opening time is under way. This will require slight changes in our relay system. The time-delay balanced-current relays will be replaced with high-speed balanced-current relays, and in some places the carrier will be speeded up slightly by using higher speed directional-relays. Another program of speeding up all 66-kv switches is being started and when that is done, changes in relays will be made to keep up with the progress.

Relays and relay systems using an electric circuit for preventing tripping has been avoided. The simple rule used is that a relay must act to complete a circuit to cause tripping, and it must be activated by a definite minimum fault current. The simple tripping circuit of Figure 2 holds true in our carrier schemes. It is very easy to have something happen to a holding circuit thus allowing the tripping of unnecessary switches. This perhaps is the main reason for not using distance relays. In several places where these were used in interconnections with other systems, outages occured more from false tripping than by actual faults. Much has been said about tripping currents of less than full-load value. In many 66-kv faults when the current was reduced by the opening of the first switch, the arc went out before the second end of the line could relay. Recently one 66-kv line grounded several times without maintaining an arc long enough to trip any of the circuit breakers.

Two instantaneous 3-phase short circuits have occurred on the 220-kv system, both caused by an operator closing a 3-phase gang-operated grounding switch on a hot line. The first occurred before the addition of instantaneous overcurrent relays and automatic separation. This caused a severe hydroelectric system outage and a much longer steam system outage. The other was tripped out by instantaneous relays at the end closest to the station. Because it happened on a long section of line, the other end did not relay, therefore the automatic separation of the system allowed the steam to carry its load without interruption. The operator cleared the line with an apparent heavy load on it and the hydroelectric system suffered only a slight interruption. This clearly illustrates the advantages of these two devices. It might be interesting to analyze the action of the normal carrier systems that rely upon voltage directional relays for tripping under these same conditions. In these cases with the voltage ratio of 2,000 to 1, not even a trace of voltage would appear across the voltage coils of the relays. This means there is no tripping with the normal setting. If the relays were biased to operate under these conditions, then all the lines would relay, as in the previously mentioned cases, and the entire hydroelectric load would be dropped. Because these two cases are the only instantaneous 3-phase short circuits on record to date, it is apparent that not much risk is being taken in an over-all operation. Another fact that sometimes is overlooked, is that when some flash-overs do go 3-phase it is possible that the arc will go out of its own accord on the first or second slip cycle because of reduction of voltage.

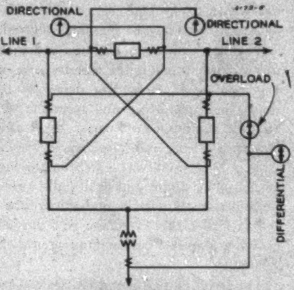


Figure 5. Station layout for one transformer bank and two lines

The separation system[4] used operates on three slip cycles, and is adjustable for two to five slip cycles. Prearranged tripping setups are made so that Big Creek hydroelectric plant has its load, Boulder hydroelectric plant has its load, and Long Beach steam plant has its load to carry when they are separated. Many tripping operations have taken place without interruption of load and with all points resynchronized in two minutes.

PROTECTION FOR EQUIPMENT

Synchronous machines generally are equipped with only the normal induction percentage differential relays for generator use. Even though generators have split windings no additional relays are used. All synchronous machines operate with neutrals ungrounded. A potential transformer connected from neutral to ground is used to initiate an alarm when faults occur.[5] In all major generating plants the unit system of generator and transformer installation is used. In some locations two generators are used with one bank of transformers. This is also true with synchronous condensers. No distribution circuits are taken directly from generator leads. The exception has occured at small unimportant generating stations. The alarm is sounded when a ground of ten per cent or more has occurred. A ground detector then is used by the operator to determine the magnitude of the ground fault. There has been much comment about using only a potential transformer for this purpose, but experience[6] has shown that it is more than justified. To date the results of operation with ungrounded neutrals have been most gratifying, and this practice has resulted in operation that has allowed use of a lower amount of spinning reserve because the chances of losing a machine have been reduced to a minimum.

J. E. Allen and S. K. Waldorf of Baltimore are to be complimented on their very valuable contribution to this subject by their work.[7] The Edison system has had all forms of ground failures since the use of ungrounded neutrals and concurs in their findings. At the present time nothing is available to alter the operating setup.

The advantages of this method of operation can be explained best by two actual cases. One midafternoon, ground alarms occurred indicating trouble on a 60,000-kva steam unit. Because the unit was required through the evening peak, it was left in service. During this time visual inspection was made of as much of the installation as possible. It was discovered that a bus support-insulator had failed. All the necessary equipment for repair was gathered and as soon as the machine could be released the repair was made in short order. If this machine had been operated with solidly grounded neutral, a heavy fault would have occurred doing much damage and the machine would have been tripped off immediately. It would have been difficult to pick up its load immediately from other sources and, of course, the unit would not have been available for that evening's peak.

In the other case, an 80 per cent ground occurred on a 45,000 kw hydroelectric generator. It was taken off the line as soon as it could be spared. Upon test, one phase was found to be solidly grounded. This 80 per cent-indicated ground showed the fault to be about 80 per cent from the neutral to the terminal. Upon removing the faulted coil a burned spot the size of a silver dollar between coils was discovered. The lower coil showed some damage too, but had not failed yet. The coils were replaced. If this machine had had a solidly grounded neutral it would have been damaged seriously. Also the instantaneous dropping of that amount of generating capacity would have caused system trouble. Many other gratifying cases could be listed.

As a result of this satisfactory record of operation a study has been started regarding new installations without direct generator protection and with only the over-all transformer and generator differential protection. By doing this, all current transformers are eliminated from generator leads, all being connected on the neutral end of the generator. The question is, do the current transformers present more of a hazard on the leads than the value gained by using the more sensitive generator protection?

Transformers are protected by induction percentage-differential relays with good results. It is pleasing to note that the manufacturers are doing something about the weakest point in transformers. Most faults on recent transformers have occurred in high-voltage bushings. Bushings have been redesigned completely for transformers for 69 kv use and above.

Protection of Distribution

The normal inverse-time-delay relays, both plain overload and directional overload, are used for phase tripping. The most difficult faults to isolate are ground faults. There is and perhaps always will be a discussion on merits of ungrounded and grounded systems. The Edison system has many of both kind. Those operating very satisfactorily ungrounded use a ground detector[8] for giving an alarm and a grounded feeder-selector[9] for indication of a faulted feeder. This method certainly gives the maximum service to customers with equal facilities, because a fault can be run down without de-energizing the line.

The grounded-neutral system has its disadvantages in that the fault-to-ground resistance[4] varies so greatly that it is impossible to have a relay sensitive enough for all conditions. This fact led to the development of sensitive ground relaying.[10] This scheme permits tripping on one per cent of current transformer rating; that is, with 100 to 5 wound current transformers, the line can be tripped with one ampere primary ground current. The maximum ground current is limited to approximately ten times minimum trip current. The two requirements for this sensitive ground protection are for when the lines run in hazardous territory and for when higher voltage circuits are used jointly with communication circuits. In the latter case tests have been made showing that upon contact, lines will trip out without even blowing subscribers' fuses. Where power and telephone utilities are serving the same customers, much is gained in taking advantage of the economy of joint use of pole lines. Hundreds of circuits are equipped with sensitive ground relaying and the almost perfect results have justified the small extra cost of installation.

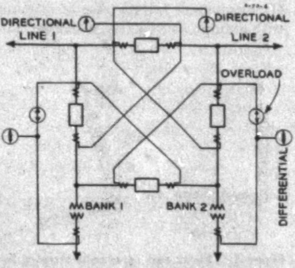


Figure 6. Station layout for two transformer banks and two lines

Station Layouts

As previously stated, the protection group has had considerable influence on station designs and layouts. Figure 5 shows the switch arrangement for a station with two lines and one transformer bank, and Figure 6 shows the switch arrangement for a station with two lines and two transformer banks. It can be seen readily that from a protection standpoint with overlapping current transformers all electrical parts are protected without the addition of bus differential protection, thus simplifying the over-all protection problem.

If five circuits are required and two are alike, such as shown for Eagle Rock substation in Figure 1 where three lines and two transformer banks are used, the six-switch setup is used. Here the bus is included in the transformer differential. For Boulder, two lines and three transformer banks are shown, and the bus is included in the line protection. With use of suspension busses a very economical installation is possible. The suspension bus requires a minimum of insulator strings, thus cutting failure hazards to a minimum.

In many station layouts long current transformer leads have been necessary. The burden of these leads has been excessive. To prevent such losses in the leads one-ampere secondary transformers are used.[11] There is a 25 to 1 gain using the same leads. The smallest size conductors from a mechanical standpoint are used which permits a considerable saving in copper and yet there is very little lead burden.

Relay Testing

All types of facilities have been tried for the testing of relays. As early as 1925 a plug-in relay was developed and studies were made to determine its value. The relay was all right, but the problems encountered in its application were more than could justify its complication. The most satisfactory method and the one that is standard in all installations in Edison stations is very reliable. On all important lines which are required to be kept in service while the relays are being tested two sets of relays with a roto-switch for switching from one group to another are installed. Where lines can be de-energized one set of relays is used with a roto-switch for testing. No lines are left in service without protection. In many installations, the second group of relays is identical to the first group; in other installations, the second group has somewhat fewer but adequate protective features. For machine differential protection the relays are inspected and adjusted during overhaul periods, but in some cases it has been justifiable to put in an alternate set of relays for machines which operate for long periods.

Electromechanical Transient Performance of Induction Motors

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Synopsis: This paper is the result of a differential analyzer study of the starting characteristics of 2-phase induction motors. The prototype of the motors investigated is the small 2-phase machine commonly used in closed cycle control systems. The performance equations of the induction motor are solved, and summary curves based upon the conclusions reached are presented in a form useful to motor designers. No published data of this type have come to the attention of the authors. It is hoped that the results herein reported can be utilized in the design of better motors for servo applications.

EARLY in the course of development of a new closed cycle control system for a-c motors, it became apparent to the authors that there was not sufficient information available about the transient performance of induction motors. High speed performance of a servomechanical system requires that the drive motor come up to required speed in as short a time as possible. At the same time other desirable features of motor design should not be subordinated.

The performance of an induction motor can be represented adequately by the set of equations derived by H. C. Stanley.[1] Since the publication of his results, relatively little work has been done on the evaluation and solution of his equations. This failure has been mainly because of the mathematical complexity which is involved in the solution of a set of simultaneous nonlinear differential equations subject to arbitary boundary conditions.

It is the purpose of this paper to present some solutions of the Stanley equations for the case of suddenly applied voltages. Previous investigators solved these equations for the case of a motor with infinite inertia,[2] a somewhat unphysical case; or they were concerned primarily with the effects of mechanical transients;[3] or they used constants typical of large motors.[4]

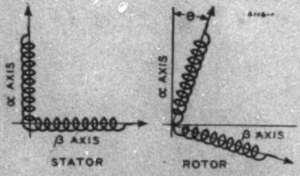

Figure 1. Stator and rotor coils showing instantaneous angular position

The motors studied have their prototype in the small 2-phase squirrel cage induction motor used in low power servomechanisms, in particular, approximately the 4-watt size. In the analysis this theoretical motor is started from rest by applying voltage to the two phases, the rotor being accelerated to synchronous speed in the case where there is no external or friction load or to some steady state speed in the case of an external or friction load.

The differential analyzer plotted the rotor speed, stator and rotor currents, and the applied voltage as functions of time. Machine parameters were varied and the results studied to determine design parameters for a motor which could start and stop in an extremely short time interval. This characteristic is necessary for successful application to a closed cycle control system.

Experimental verification for one of the motors investigated has been obtained to a degree which makes valid the broad conclusions drawn from the remainder of the work.

Assumptions

In accordance with the limitations of the performance equations for the induction motor, the following simplifying assumptions were made in this analysis:

1. Both rotor and stator have symmetrical windings.
2. The coefficient of mutual inductance is

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The authors wish to acknowledge the assistance of the differential analyzer laboratory of the Moore School and of Chuan Chu, who assisted with the experimental work. This paper is part of the results obtained in work under contract between the Bureau of Ordnance, Navy Department, and the University of Pennsylvania.

All relays are given careful inspection by test engineers. The 220-kv line protection is given a routine check twice a year and others are inspected for periods varying from one complete secondary and tripping test a year to one every other year.

The two group relay setups have proved valuable many times when system operating conditions changed. At such times one group of relays was set for one condition of operation while the other was set for another.

All important switches are given a trip test by the operators each month according to a prearranged schedule. With all this precaution and continual checkup the comparatively simple relaying system of the Edison Company has performed most satisfactorily.

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a cosinusoidal function of the electrical angle between the rotor and stator axes.

3. The rotor is smooth, and the self-inductances of all windings are independent of rotor position.

4. The effects of saturation, hysteresis, and eddy currents in the iron circuits are neglected.

5. The machine impedances are independent of rotor speed.

6. The motor is connected to an infinite bus through zero impedance.

7. All electrical switching is accomplished in zero time.

8. The stator and rotor neutrals are ungrounded.

Conclusions

1. The performance of an induction motor may be specified completely by five parameters.

τ_s = the stator time constant
τ_r = the rotor time constant
τ_m = the mechanical time constant
k = the coefficient of coupling
T_L = the load torque

2. The mechanical time constant, τ_m, should be small for rapid acceleration of the rotor. Methods for making τ_m small are discussed in a subsequent paragraph.

3. The stator and rotor time constants have optimum values for the rise time to be a minimum. The ratio of the optimum time constants is approximately unity.

4. The rise time is independent of shaft loads which are proportional to the velocity.

5. A small air gap reduces the rise time, but a point of diminishing returns can be reached between further decreases in air gap and decreases in rise time.

6. Starting an induction motor by applying voltage to one phase when the other phase has been previously energized yields the same rise time as if both phases were energized at the same time, within the range of parameters investigated.

Units Used in Analysis

In the analysis of problems involving many parameters it is usually convenient to rewrite the equations in terms of dimensionless variables. This is done commonly in electric machinery analysis, and the equations so derived are said to be in a per unit system. Common kilovolt-ampere and time bases are chosen in the usual per unit system, and voltage bases are chosen for each part of the system. In this paper it was more convenient to choose as a common voltage base the voltage actually applied to the machine and separate impedance bases for the rotor and stator. Base time was chosen as the time in seconds for the revolving field to move 2π electrical radians as in the usual per unit system. Per unit values as used in this paper mean the actual values divided by the appropriate base values. Thus, per unit current means the ratio of the actual current in amperes to the current that would be obtained if the voltage actually applied to the machine were applied to an impedance equal to the base value of impedance. Therefore, base values are not necessarily coincident with name plate ratings of the machine.

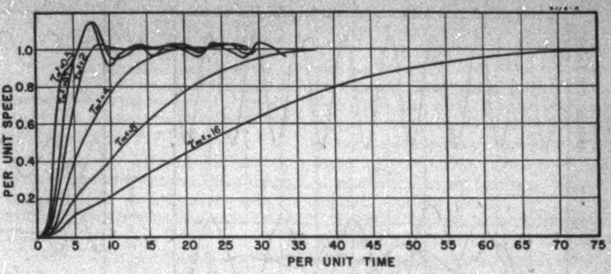


Figure 2. Effect of varying τ_m

$\tau_s=4$ $k=0.75$
$\tau_r=2$ $T_L=0$

Results From the Differential Analyzer

The differential analyzer plotted directly the stator current, the α and β components of rotor currents, and the rotor speed as functions of time. Hereafter, the α and β components of rotor current will be called simply rotor currents. Typical curves are shown in Figures 2 through 8.

Effect of Mechanical Time Constant

Figure 2 indicates clearly that when the mechanical time constant τ_m is decreased, there is a corresponding decrease in the rise time. For purpose of definition, rise time is taken to mean the time required to reach 85 per cent of full steady state speed. In the case of no load, this corresponds to a slip of 15 per cent. There does not seem to be any uniformity on definition of rise time for motors. In fact, there is not much agreement among workers in closely allied fields, the general opinion being to define rise time to best suit a particular problem. The square of the mechanical time constant for the 2-phase machine is defined as

$$\tau_m^2=\frac{\omega^4 JL}{2q^2E^2k^2}$$

It is seen to be a function of six different quantities. Of these, only five can be varied at will without changing the other significant machine constants. In particular, the curves shown can represent linear changes in the rotor inertia or the stator inductance; or the curves can show changes inversely proportional to the square of the product of applied voltage, number of pairs of poles, and coefficient of coupling. The curves are sufficiently general to admit interpretation in either direction.

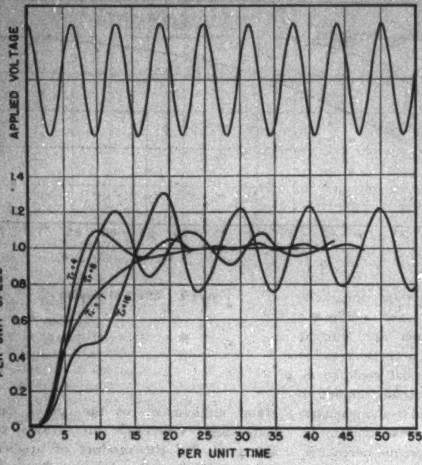
It can be seen from these curves that an induction motor at normal voltage does not rise to speed in a smooth fashion. This can be explained by the fact that during the starting period there exists a damped oscillation in the electric torque.

Effect of Rotor and Stator Time Constants

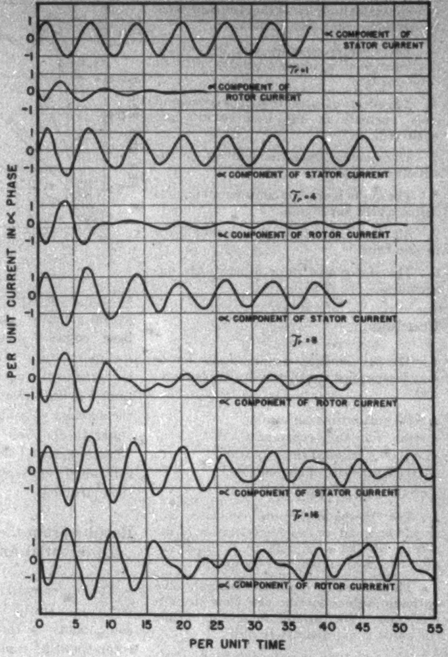
The magnitude and duration of the induced voltage and current in the rotor circuit would be obviously a function of the rotor parameters. In the case of a squirrel cage rotor it is virtually impossible to measure inductance and resistance. However, it is possible by a series of physical measurements (Appendix III) to compute a number which can represent the ratio of the rotor inductance to the rotor resistance. The product of this ratio and the angular frequency is termed the rotor time constant and is given by the relationship $\tau_r=\omega l/r$.

Figure 3 illustrates the effect of changes in this time constant, keeping other machine parameters fixed. It is seen that there is an optimum value for the rotor time constant in so far as rapid rise time is concerned. The rather large oscillations in speed, not at fundamental frequency, are oscillations at the natural period of the system, determined by the inertia and the electric torque acting as a spring.

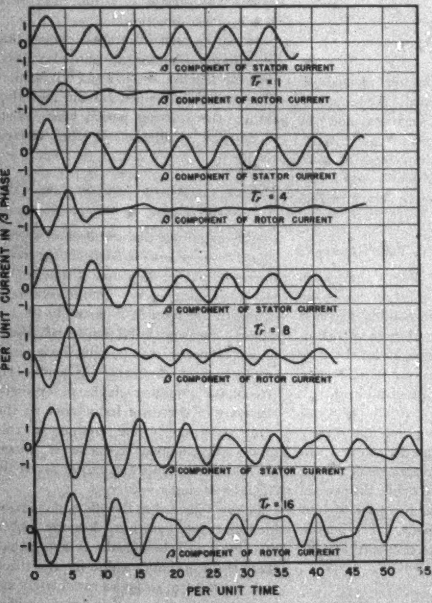
This figure also shows the variations of the rotor and stator currents as functions



(a)



(b)



(c)

Figure 3. Effect of varying τ_r

$\tau_m{}^2=2.5$ $k=0.75$
$\tau_s=4.0$ $T_L=0$

(a). Per unit speed and applied voltage versus per unit time

(b). Per unit current in α phase versus per unit time

(c). Per unit current in β phase versus per unit time

of time. It is seen that when the rotor currents have decreased to a small value, the rotor has gone up to synchronous speed. The current surges in the rotor rarely last for more than one cycle. Therefore, little attention need be given to its peak heat dissipating properties.

The stator inductance and resistance combine in an analogous manner to form the stator time constant $\tau_s=\omega L/R$. Figure 4 shows the motor speed as a function of time. It can be observed that there is an optimum value of τ_s for this machine, other constants being equal.

Effect of Friction Load

In the majority of applications of servo motors used in low power controls, there is little or no load on the motor. The curves previously discussed were all for the case of no load. In certain other applications, however, a damping load proportional to velocity is of importance. The effect of a load on the performance of an actual motor is illustrated by the speed response curves shown in Figure 5. The time to reach steady speed is almost independent of the load. Of course, the final speed is dependent upon load. It is interesting to see from this curve that for most loads encountered in servo work, even up to a friction torque coefficient equal to the base torque of the machine, there is no appreciable change in the speed–time curve. It is only for the much higher values of load that there is a decided drop in speed.

Effects of Switching

A servomechanism must have rapid response in accordance with variations in input signal. Consequently, a servo motor must be able not only to rise to a steady state speed rapidly, but also to reverse in a short time. An investigation was made of the reversing or switching of a motor, with possible application to plugging a motor to zero speed. Typical results are shown by the solid lines in Figure 6, where the variable parameter is τ_m. The parameters used in this set of curves are those of an actual motor. It is seen that the lower the value of τ_m, the shorter can be the reversing time. Here again is observed the decided oscillation in the motor speed after the reversal. In this study, reversal of the motor was accomplished by shifting instantaneously the phase angle of the voltage applied to one phase of the motor by 180°. The effect of this switching on the currents can be observed clearly.

The β phase was switched in this investigation. It is seen that the rotor and stator currents reverse in phase and go through a transient period, in which the magnitude of the transient current may be several times the normal steady state value. During this transient period, the α phase currents likewise change, due to induced effects, but here there is a change of magnitude only, accompanied by a time lag due to the electrical constants of the machine. The initially large rotor currents result in a large value of electric torque in a reverse direction; hence, the motor speed starts to drop almost instantaneously. The acceleration during this period never can be infinite because of circuit time lags, but it can be extremely large. For example, the average acceleration of the steepest speed curve shown in Figure 6 is approximately 1. This means that the motor comes to rest from full speed in one direction in an interval of time of 1 unit, which for a 60-cycle voltage is 1/377 second or 2.65 milliseconds. In terms of cycles, this is 0.159 cycle or about 1/6 cycle. Response of this order of magnitude is desirable for a rapid servomechanism.

SEQUENTIAL ENERGIZING OF PHASES

Most applications of servomechanisms which use a 2-phase induction motor for the output are operated with one phase permanently connected to a source, the second phase responding to a signal. Figure 7 proves that better over-all operation of the motor may be achieved by connecting both phases simultaneously. In Figure 7, the initial conditions were as follows:

1. Both phases were connected at the same time, and the motor accelerated up to synchronous speed.
2. With the motor at rest, the α phase was energized, and the β phase was short-circuited. This latter condition was continued for three cycles of the applied voltage. Then full voltage was applied to the β phase. The speed response of the motor as a result of sequential application of voltage, was exactly the same as if both voltages were applied at the same time.

As a result, it would seem desirable not to operate a motor with one phase always energized, if for no other reason than to keep the motor from heating when not in useful operation.

In order not to confine this study to a small motor, a set of curves is shown in Figure 8 representative of a much larger machine. For the constants chosen, it is seen that the motor accelerates unevenly, and when the average steady state is reached an oscillation in speed at approximately fundamental frequency occurs. Presumably, had the solution been carried further, these oscillations would have died out. The rotor and stator currents are particularly interesting for this set of parameters. The transients are prolonged and of higher value than in the motors previously considered. A decided d-c component in the rotor and stator circuits is seen to be present.

Analysis of Differential Analyzer Results

The general results of this study may be summarized in a set of curves showing the over-all effects of varying the time constants and physical dimensions of the induction motor. Typical summary sheets are shown in Figures 9 through 14.

The time for the experimental motor to accelerate from rest to 85 per cent of its steady state speed and the time for the current to decay to steady state value are plotted in Figure 9 as functions of the square of the mechanical time constant τ_m. After steady state was reached, for each of the values of τ_m the voltage applied to one phase was reversed, resulting in a reversal of the rotor rotation. This reversal introduced a transient current. The time curves for the speed to change to 85 per cent speed in the reverse direction are shown as a dotted line in Figure 9. The broken lines are drawn to represent a motor having a slightly different value of τ_s and τ_r.

The effect of a low value for the mechanical time constant can be obtained by increasing the applied voltage above normal. But in a machine this variation is limited by insulation difficulties and by a saturation, an effect wholly neglected in the present study.

The coefficient of coupling k cannot be changed to any great extent, even by machine design, but the number of pairs of poles q can be changed. Since this factor does not appear in any other machine parameter, it is reasonable to say that increasing the number of pairs of poles to a limit bounded by the physical size of the machine is to be preferred to increasing the voltage.

Of course, when the number of poles is increased, the synchronous speed is decreased and the torque increased in direct proportion. Servo motors usually drive a load through a step-down gear ratio, so that the output speed can be kept the same by reducing this ratio. This results in more efficient over-all operation. When the number of poles is increased, the rotor inertia probably must be increased also. If the ratio of the new moment of inertia to the original moment of inertia is less than the square of the increase in the number of pairs of poles, τ_m will be reduced.

The theoretical results obtained on the

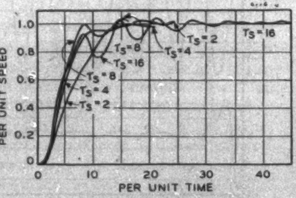


Figure 4 (above). Effect of varying τ_s

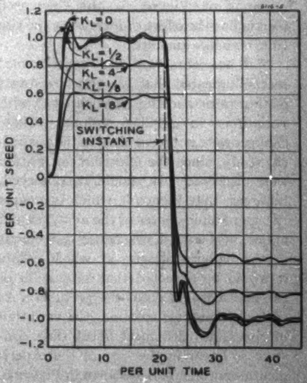
$\tau_m^2=2.5$ $k=0.75$

$\tau_r=2.0$ $T_L=0$

Figure 5 (right). Effect of varying K_L

$\tau_s=2.35$ $\tau_m^2=0.356$

$\tau_r=3.50$ $k=0.815$



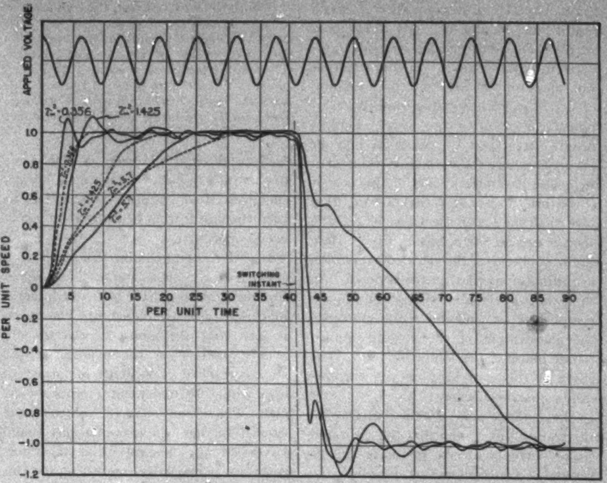


Figure 6a. Comparison of theoretical and experimental curves for various τ_m's

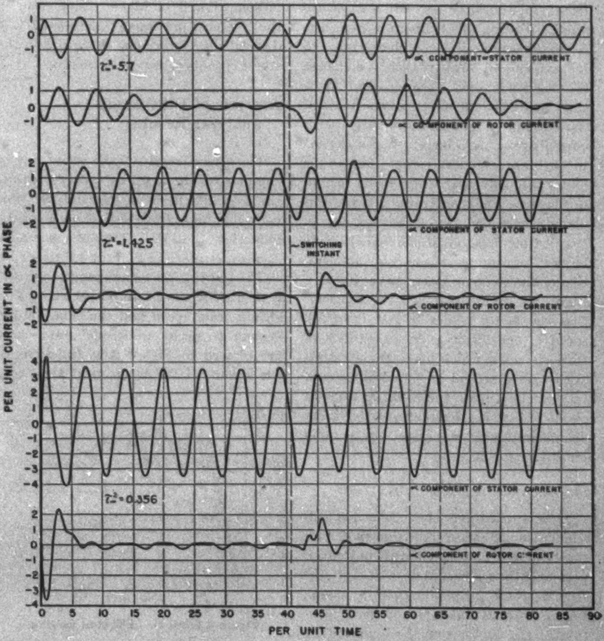
$\tau_s=2.35$ $k=0.815$
$\tau_r=3.50$ $T_L=0$

Per unit speed and applied voltage versus per unit time. The dotted curves were determined experimentally

differential analyzer were verified experimentally in the case of variations in applied voltage or τ_m. The parameters of an actual motor were used on the differential analyzer, the results of which are shown in Figure 6. For an experimental check, this motor was shaft coupled to an a-c rate generator with one phase of the generator excited with 1,000 cycles. The voltage induced in the second phase was proportional to the speed of rotation of the motor rotor. The generator output was fed to a cathode ray oscilloscope, and a photographic record was made of this voltage as the motor accelerated from standstill. Measurements were made for applied voltages of one, two, and four times normal. The envelopes of the resulting curves are plotted on Figure 6 as a dashed line. It is seen that at voltages above normal there is better agreement between the theoretical and experimental result. For all voltages, the time to rise to full speed is approximately the same for the theoretical and experimental tests, agreement being within 30 per cent at full speed.

Some differences exist between the motor theoretically studied and that used in the experimental setup. The added inertia due to the rate generator rotor increases the effective inertia of the motor by 13.3 per cent. The effect of this increased parameter increases the square of the mechanical time constant by a like amount. As seen from the summary curve *d* of Figure 9, this increased inertia has the effect of increasing the rise time by a like amount.

Measurements on the experimental motor indicated that the time constants of the stator and the rotor changed with variations in applied voltage. These changes were not recognized in the analytical study, since the effects of saturation were neglected. It should be observed, however, that Figure 6, which compares the speed-time curves of the analyzer and actual systems, shows better agreement at the higher voltages, at which points it would be expected that saturation of the motor would enter the problem. As with the analyzer solution, the variation of rise time with mechanical time constant is linear in the case of the experimental motor. This is shown in Figure 9.



At first thought, it would be desirable that the rotor time constant be made as small as possible. The argument for this small value is that the rotor currents would die out rapidly, and, therefore,

Figure 6b. Comparison of theoretical and experimental curves for various τ_m's

$\tau_s=2.35$ $k=0.815$
$\tau_r=3.50$ $T_L=0$

Per unit current in α phase versus per unit time the motor at no load would reach synchronous speed in a short time. Inspection of the summary curve in Figure 10 shows that this is not the case. For a fixed set of machine parameters, other than the rotor time constant, it is seen that there is an optimum value to the rotor time constant, which will result in a minimum rise time. This value is not defined well. For the machine parameters whose curve is shown, the range of time constants would be from two to six. From other observed data, a general rule would be that $\tau_r < \tau_s < 3\tau_r$ for short rise times. The rotor time constant, $\tau_r = \omega l/r$, is rather difficult to ascertain theoretically in the design of a rotor, because in the case of a squirrel cage machine the effective values of the self-inductance and rotor resistance are functions of the applied voltage. Of course, a wound rotor can be designed to fit any desired design parameters. The low values of τ_r desired can be interpreted either as due to a low value of self-inductance or to a large value of rotor resistance.

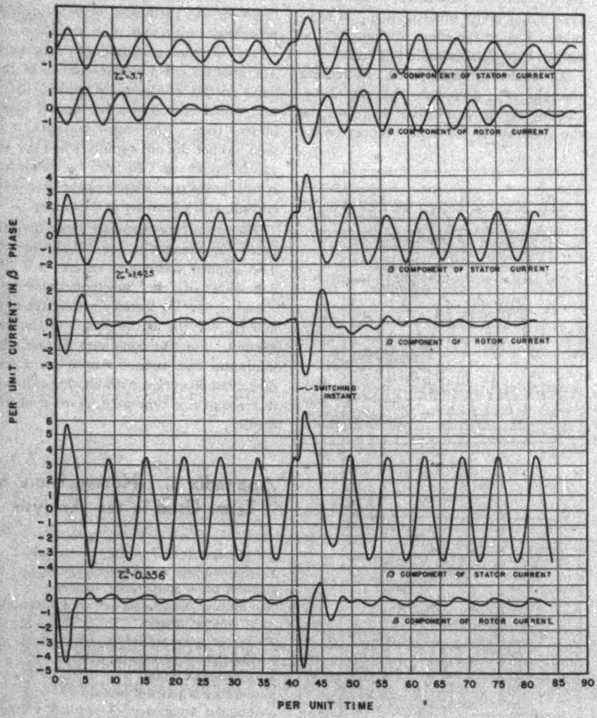


Figure 6c. Comparison of theoretical and experimental curves for various τ_m's

$\tau_s = 2.35$ $\qquad k = 0.815$

$\tau_r = 3.50$ $\qquad T_L = 0$

Per unit current in B phase versus per unit time

The effect of low rotor time constant is reflected back to the stator, as can be seen in the curve showing the time for the current to reach its steady state value. The low value of τ_r or the high value of rotor resistance is transferred to the stator, which then has an increased effective resistance or a decreased time constant. This in turn tends to shorten the period of the transient stator current.

The rise time of the induction motor is a function of the stator time constant in much the same way as it is a function of the rotor time constant; namely, in the presence of an optimum value of stator time constant there is a minimum rise time. For the curve shown in Figure 11, $\tau_s \cong 3\tau_r$. This is in the range of τ_s, as indicated from the τ_r variation curve. Although the rise time increases with further decrease in stator time constant, the actual time for the stator current to reach its steady state value decreases with a decrease in stator time constant. This is understandable, because a low value of τ_s may be interpreted as an increased stator resistance, and hence a shorter transient period.

Figure 12 shows some rather surprising results. This summary curve shows over-all results for changes in friction load torque coefficient. The product of friction load torque coefficient K_L and speed is the total damping torque. It is seen that the motor reaches 85 per cent of its final speed in a time independent of the load torque. Furthermore, the time for the current to decay to a steady state value is also independent of load conditions. The final speed obtained by the motor for each load is shown. It is seen that even for a load torque coefficient eight times base value the final speed is as high as 55 per cent of synchronous speed.

It is apparent that changing the air gap will alter the magnetic structure in such a way as to reflect this change upon the parameters of the stator and the rotor. Although the inductance coefficients are related to the coefficient of coupling by the relationship $k = M/\sqrt{Ll}$, it is not sufficient to change only k in considering changes in air gap. That this is so may be seen from the following rather approximate analysis.

The magnetic circuit for both the stator and the rotor may be thought of as being composed of two flux paths, one through the iron structure itself, mutual to the stator and the rotor, designated by a permeance P_m, and the other a flux leakage path, different for the stator and the rotor, designated by permeances P_{ls} and P_{lr} for the stator and rotor, respectively. On the further assumption that the inductance coefficients are proportional to the square of a factor N, the number of turns, there can be written

$$L = N_s^2(P_m + P_{ls}) \tag{1}$$

$$l = N_r^2(P_m + P_{lr}) \tag{2}$$

$$M = N_s N_r P_m \tag{3}$$

The coefficient of coupling then becomes

$$k = \frac{M}{\sqrt{Ll}}$$

$$= \frac{P_m}{\sqrt{(P_m + P_{ls})(P_m + P_{lr})}} \tag{4}$$

If $P_{ls} = P_{lr} = P_l$, a not too approximate

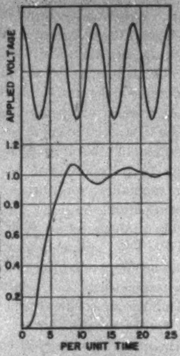


Figure 7a. Per unit speed and applied voltage versus per unit time

$\tau_s = 7.00$ $k = 0.857$
$\tau_r = 3.50$ $T_L = 0$
$\tau_m^2 = 3.85$

assumption, the coupling becomes

$$k = \frac{P_m}{P_m + P_l}$$

$$= \left(1 + \frac{P_l}{P_m}\right)^{-1} \tag{5}$$

$$k \approx 1 - \frac{P_l}{P_m} + \left(\frac{P_l}{P_m}\right)^2 \ldots\ldots \tag{6}$$

This term is seen to be dependent upon the ratio of the permeance to leakage flux to the permeance to mutual flux. For most cases it is sufficient to consider only the first two terms. The self-inductance of the stator and rotor now can be written

$$L = N_s^2 \frac{P_m}{k} \tag{7}$$

$$l = N_r^2 \frac{P_m}{k} \tag{8}$$

The mutual permeance, being dependent upon the magnetic path through the rotor and the stator, must include the effect of the air gap. The mutual permeance can then be written as

$$P_m = \frac{1}{\frac{1}{P_l} + \frac{1}{P_{gap}}}$$

$$= \frac{P_l P_{gap}}{P_l + P_{gap}} \tag{9}$$

and, because the permeance is proportional to the factor μ/l, where μ is the permeability and l is the effective length of the magnetic path, it is seen that to a first approximation, $P_l >> P_{gap}$, and

$$P_m = P_{gap} \propto \frac{l}{l_{gap}} \tag{10}$$

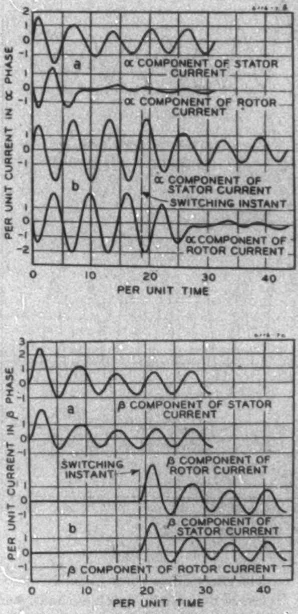
Therefore, from equations 7 and 8, we have

$$N_s^2 = kLl_{gap} \tag{11}$$

$$N_r^2 = kll_{gap} \tag{12}$$

These relationships were used to vary theoretically the parameters of the prototype motor under investigation. A summary curve of the effect of changing the air gap spacing is shown in Figure 13.

As would be anticipated, decreasing the air gap decreases the rise time. Cutting the air gap in half decreases the rise time by 21 per cent. Since the air gap in the prototype motor as presently manufactured is greater than required from physical considerations, it would seem desirable to revise the design of such motors to decrease the air gap and so decrease the rise time. There is, however, a point of diminishing returns. If the air gap were made infinitely small, even zero, there would be a finite rise time, for zero gap would, on our assumptions, result in the coefficient of coupling approaching unity and in a complete revision of the performance equations. The greatest effect would seem to be the finite inertia of the rotor, which, in itself, would not allow a rise to full speed in zero time. The stator currents, it is seen, do not die out rapidly when the air gap is decreased. In fact, the tendency is to increase the length of the transient current with a decrease in the air gap.



Figures 7b and 7c. Per unit current versus per unit time, showing effect of energizing one or both phases

(b). In α phase
(c). In β phase

Curve *a* is for the case when both phases were energized at the same time

For curve *b*, the α phase was energized for three cycles of the applied voltage with the β phase short circuited. Then the β phase was energized

$\tau_s = 7.00$ $k = 0.857$
$\tau_r = 3.50$ $T_L = 0$
$\tau_m^2 = 3.85$

The prototype machine was investigated for changes in applied frequency. The applied frequency enters into the definition of the machine constants τ_r and τ_s, varying linearly with frequency, and τ_m, varying as the square of the frequency. For the constants chosen, the minimum rise time occurred at a frequency of 60 cycles, with an almost linear increase in rise time with increase in frequency.

Appendix I. Nomenclature of Terms Used in the Analysis

E = voltage applied to the motor, volts
L = stator phase inductance, henrys
l = rotor phase inductance, henrys
M = mutual inductance between rotor and stator, henrys
R = stator phase resistance, ohms
r = rotor phase resistance, ohms
p = number of phases
q = number of pairs of poles
ω = angular frequency of applied voltage, radians per second
J = rotor moment of inertia, watt seconds cubed
t' = actual time, seconds
θ_m = rotor position with respect to the stator, mechanical degrees
θ = rotor position with respect to the stator, electrical degrees
T_L' = actual total load torque, watt seconds
I_α', I_β' = actual stator currents in positive α and β axes, respectively, amperes
i_α', i_β' = actual rotor currents in positive α and β axes, respectively, amperes
I_α', I_β = per unit stator current in positive α and β axes, respectively
i_α', i_β = per unit rotor current in positive α and β axes, respectively
T_L = per unit load torque
K_L = load torque coefficient; $T_L = K_L \frac{d\theta}{dt}$

$\tau_s = \frac{\omega L}{R}$ = stator time constant

$\tau_r = \frac{\omega l}{r}$ = rotor time constant

$\tau_m = \omega \left(\frac{J}{qT_n}\right)^{1/2}$ = mechanical time constant

$T_n = \frac{pqE^2k^2}{\omega^2 L}$ = base torque, watt seconds

$I_n = \frac{E}{\omega L}$ = base stator current, amperes

$i_n = \frac{EM}{\omega Ll}$ = base rotor current, amperes

$t = \omega t'$ = base time

Appendix II. Derivation of Equations Solved on the Differential Analyzer

The Stanley equations for the operation of a 2-phase induction motor can be written as follows:

$$RI_\alpha' + L\frac{dI_\alpha'}{dt'} + M\frac{di_\alpha'}{dt'} = E \cos \omega t' \tag{13}$$

$$RI_\beta' + L\frac{dI_\beta'}{dt'} + M\frac{di_\beta'}{dt'} = E \sin \omega t' \tag{14}$$

$$M\frac{dI_\alpha'}{dt'} + ri_\alpha' + l\frac{di'_\alpha}{dt'} + (MI_\beta' + li_\beta')\frac{d\theta}{dt'} = 0 \tag{15}$$

$$M\frac{dI_\beta'}{dt'} + ri_\beta' + l\frac{di_\beta'}{dt'} - (MI_\alpha' + li\alpha')\frac{d\theta}{dt'} = 0 \tag{16}$$

$$J\frac{d^2\theta_m}{dt'^2} = \frac{qp}{2}M(I_\beta' i_\alpha' - I\alpha' i_\beta') - T_L' \tag{17}$$

Define now a base stator current, $I_n = E/\omega L$, and a base rotor current, $i_n = EM/\omega Ll$. Then, dividing equations 13 and 14 by these base values, there results

$$\frac{R}{\omega L}I_\alpha + \frac{dI_\alpha}{dt} + \frac{M^2}{Ll}\frac{di_\alpha}{dt} = \cos t \tag{18}$$

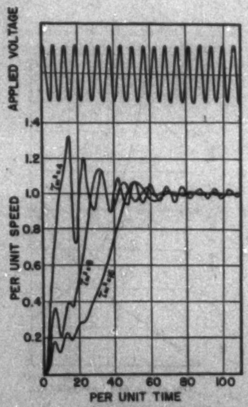


Figure 8a. Per unit speed and applied voltage versus per unit time

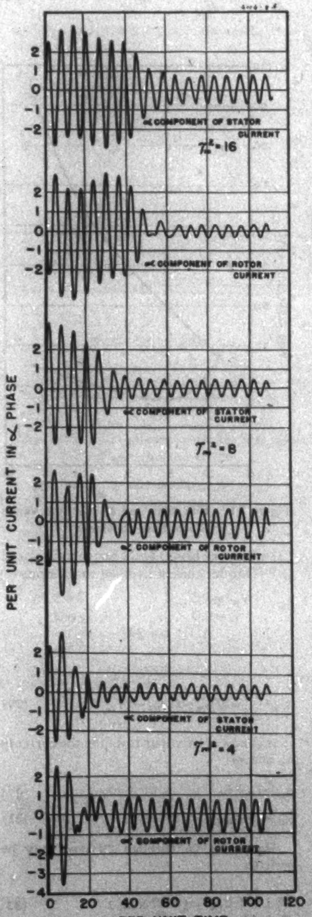
$\tau_s = 32$ $k = 0.815$
$\tau_r = 16$ $T_L = 0$

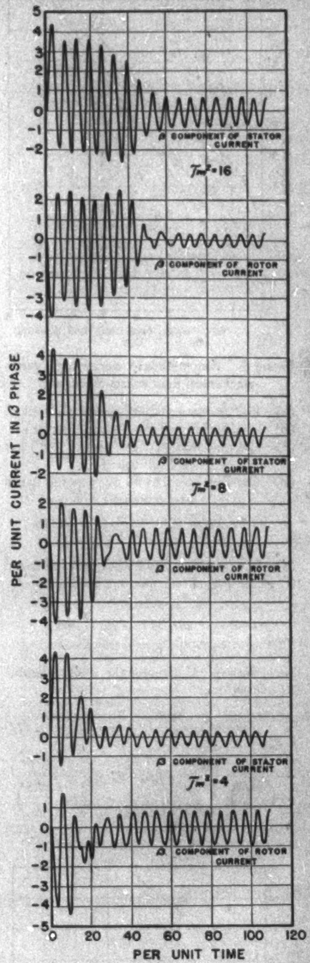
Figure 8b (below). Per unit current in α phase versus per unit time

$\tau_s = 32$ $k = 0.815$
$\tau_r = 16$ $T_L = 0$

Figure 8c (right). Per unit current in β phase versus per unit time

$\tau_s = 32$ $k = 0.815$
$\tau_r = 16$ $T_L = 0$





$$\frac{R}{\omega L}I_\beta + \frac{dI_\beta}{dt} + \frac{M^2}{Ll}\frac{di_\beta}{dt} = \sin t \tag{19}$$

In a similar manner equations 15 and 16 transform to

$$\frac{dI_\alpha}{dt} + \frac{r}{\omega l}i_\alpha + \frac{di_\alpha}{dt} + (I_\beta + i_\beta)\frac{d\theta}{dt} = 0 \tag{20}$$

$$\frac{dI_\beta}{dt} + \frac{r}{\omega l}i_\beta + \frac{di_\beta}{dt} - (I_\alpha + i_\alpha)\frac{d\theta}{dt} = 0 \tag{21}$$

The developed electric torque and the load torque are balanced by the product of the moment of inertia of the rotor and its

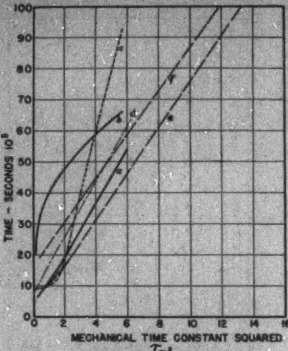


Figure 9. Rise time and decay time versus mechanical time constant squared

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear
(*c*)—Rise time from 100 per cent to −85 per cent full speed
(*d*)—Rise time from 0 to 85 per cent full speed (experimental)

$\tau_s=2.35$ $k=0.815$
$\tau_r=3.50$ $\omega=377$ $T_L=0$

(*e*)—Rise time from 0 to 85 per cent full speed
(*f*)—Time for current transient to disappear

$\tau_s=4$ $k=0.75$
$\tau_r=2$ $\omega=377$ $T_L=0$

acceleration. Hence, for the whole machine (p phases),

$$\frac{J}{q}\frac{d^2\theta}{dt'^2}=\frac{pq}{2}M(I_\beta'i_\alpha'-I_\alpha'i_\beta')-T_L' \qquad (22)$$

$$\frac{\omega^2Jl}{pq^2\left(\frac{ME}{\omega L}\right)^2}\frac{d^2\theta}{dt^2}=1/2(I_\beta i_\alpha-I_\alpha i_\beta)-\frac{T_L'l}{pq\left(\frac{ME}{\omega L}\right)^2} \qquad (23)$$

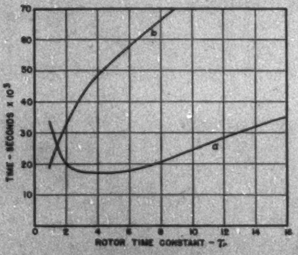


Figure 10. Rise time and decay time versus rotor time constant

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear

$\tau_m{}^2=2.5$ $k=0.75$
$\tau_s=4.0$ $T_L=0$
$\omega=377$

$$\frac{\omega^2J}{qT_m}\frac{d^2\theta}{dt^2}=1/2(I_\beta i_\alpha-I_\alpha i_\beta)-T_L \qquad (24)$$

Equations 18 through 21 and equation 24 describe the performance of the induction motor on a per unit base, in accordance with our definitions of base values. If these equations are rewritten in terms of the time constants defined in Appendix I, we have

$$\frac{I_\alpha}{\tau_s}+\frac{dI_\alpha}{dt}+k^2\frac{di_\alpha}{dt}=\cos t \qquad (25)$$

$$\frac{I_\beta}{\tau_s}+\frac{dI_\beta}{dt}+k^2\frac{di_\beta}{dt}=\sin t \qquad (26)$$

$$\frac{dI_\alpha}{dt}+\frac{i_\alpha}{\tau_r}+\frac{di_\alpha}{dt}+(I_\beta+i_\beta)\frac{d\theta}{dt}=0 \qquad (27)$$

$$\frac{dI_\beta}{dt}+\frac{i_\beta}{\tau_r}+\frac{di_\beta}{dt}-(I_\alpha+i_\alpha)\frac{d\theta}{dt}=0 \qquad (28)$$

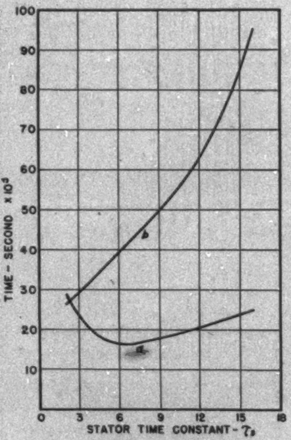


Figure 11. Rise time and decay time versus stator time constant

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear

$\tau_m{}^2=2.5$ $k=0.75$
$\tau_r=2.0$ $T_L=0$
$\omega=377$

$$2\tau_m{}^2\frac{d^2\theta}{dt^2}=(I_\beta i_\alpha-I_\alpha i_\beta)-2T_L \qquad (29)$$

It is now convenient to define the currents x_α and x_β

$$x_\alpha=I_\alpha+i_\alpha \qquad (30)$$

$$x_\beta=I_\beta+i_\beta \qquad (31)$$

Equations 25 through 29 now may be written

$$\frac{I_\alpha}{\tau_s}+\frac{dx_\alpha}{dt}+(k^2-1)\frac{di_\alpha}{dt}=\cos t \qquad (32)$$

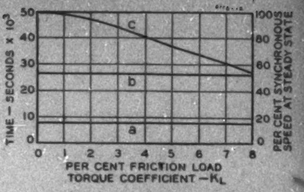


Figure 12. Rise time and decay time versus friction load torque coefficient

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear
(*c*)—Per cent synchronous speed at steady state

$\tau_m{}^2=0.356$ $k=0.815$
$\tau_s=2.35$ $\omega=377$
$\tau_r=3.50$

$$\frac{I_\beta}{\tau_s}+\frac{dx_\beta}{dt}+(k^2-1)\frac{di_\beta}{dt}=\sin t \qquad (33)$$

$$\frac{i_\alpha}{\tau_r}+\frac{dx_\alpha}{dt}+x_\beta\frac{d\theta}{dt}=0 \qquad (34)$$

$$\frac{i_\beta}{\tau_r}+\frac{dx_\beta}{dt}-x_\alpha\frac{d\theta}{dt}=0 \qquad (35)$$

$$2\tau_m{}^2\frac{d^2\theta}{dt^2}=(I_\beta i_\alpha-I_\alpha i_\beta)-2T_L \qquad (36)$$

Upon integrating once with respect to time, we have

$$\frac{1}{\tau_s}\int I_\alpha dt+x_\alpha+(k^2-1)i_\alpha=\sin t \qquad (37)$$

$$\frac{1}{\tau_s}\int I_\beta dt+x_\beta+(k^2-1)i_\beta=-\cos t \qquad (38)$$

$$\frac{1}{\tau_r}\int i_\alpha dt+x_\alpha+\int x_\beta d\theta=0 \qquad (39)$$

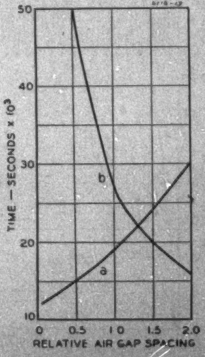


Figure 13. Rise time and decay time versus relative air gap spacing

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear

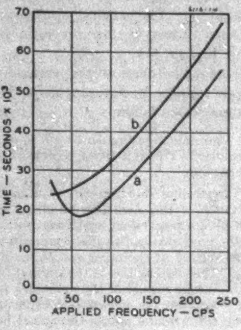


Figure 14. Rise time and decay time versus applied frequency

(*a*)—Rise time from 0 to 85 per cent full speed
(*b*)—Time for current transient to disappear

$$\frac{1}{\tau_r}\int i_\beta dt+x_\beta-\int x_\alpha d\theta=0 \tag{40}$$

$$2\tau_m{}^2\frac{d\theta}{dt}=\int I_\beta i_\alpha dt-\int I_\alpha i_\beta dt-2\int T_L dt \tag{41}$$

$$\theta=\int\frac{d\theta}{dt}dt \tag{42}$$

Equations 37 through 42 were solved on the differential analyzer, subject to various boundary conditions and various sets of parameters as discussed in the body of the paper.

Appendix III. Experimental Determination of Machine Constants

The range of parameters to be investigated was determined by making the standard blocked rotor and free rotor tests, under the assumptions that the motors had no hysteresis or iron circuit losses.

The equivalent circuits for these tests are shown in Figures 15a and b. From the first of these figures, for the blocked rotor case, the equivalent input impedance is given by

$$Z_{in}=\left(R+\frac{\omega^2M^2r}{r^2+\omega^2l^2}\right)+j\left(\omega L-\frac{\omega^2M^2l}{r^2+\omega^2l^2}\right) \tag{43}$$

If measurements of the current I_B, voltage E_B, and power P_B are now made, there result

$$\frac{P_B}{I_B{}^2}=R+\frac{\omega^2M^2r}{r^2+\omega^2l^2} \tag{44}$$

$$\sqrt{\left(\frac{E_B}{I_B}\right)^2-\left(\frac{P_B}{I_B{}^2}\right)^2}=\omega L-\frac{\omega^2M^2l}{r^2+\omega^2l^2} \tag{45}$$

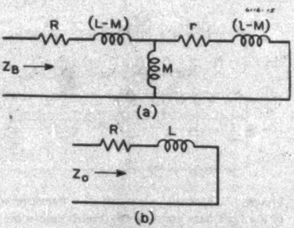


Figure 15. Equivalent circuits for one phase of an induction motor

(*a*)—Rotor blocked
(*b*)—Rotor free

For the free rotor case shown in Figure 15b, similar measurements are made, with like quantities designated by the subscript *o*. There result

$$\frac{P_o}{I_o{}^2}=R \tag{46}$$

$$\sqrt{\left(\frac{E_o}{I_o}\right)^2-\left(\frac{P_o}{I_o{}^2}\right)^2}=\omega L \tag{47}$$

Combining the results of the blocked rotor and the free rotor tests, we have the ratio $\tau_r=\omega l/r$, the rotor time constant. The coefficient of coupling can be determined from these same measurements as follows:

$$A=\frac{\omega^2M^2l}{L(r^2+\omega^2l^2)}=\frac{M^2}{Ll}\left(\frac{\omega\tau_r{}^2}{1+\tau_r{}^2}\right) \tag{48}$$

Since the quantity A can be computed from the test measurements, we have immediately that the coefficient of coupling is given by the square root of the term

$$k^2=\frac{M^2}{Ll}=A\left(\frac{1+\tau_r{}^2}{\omega\tau_r{}^2}\right) \tag{49}$$

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Oscillations of a High-Voltage Secondary Winding

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THE PHYSICAL THEORY of the characteristic oscillations of a high-voltage winding as a primary, that is, when the voltage disturbances are impressed directly across its terminals, is now well known.[1] However its behavior as a secondary with respect to the disturbances is not as well known. Although papers[2–4] have been published on this latter subject, this paper presents a physical theory believed to be novel and helpful.

The theory is of interest not merely as basic knowledge of transformer transients, but also as having applications to design. For instance, contrary to what generally is assumed, it is found that depending on its constants, the same oscillatory winding may develop higher transient voltages as a secondary than as a primary winding. This has a bearing on the design of both windings and ratio adjusters, the latter for the reason that taps from the various parts of the winding are brought together into a small space.

While the main feature of the present theory is the physical picture which it yields, it is not merely qualitative, but is fully quantitative and is formulated in terms of constants with which designers are familiar and which may be calculated or tested by methods well known to them.

General Nature of Theory

As a primary the characteristic modes of oscillation of a winding with the neutral end grounded are half waves (and their multiples). As a secondary they are neither half waves nor quarter waves but intermediate fractions of a wave. They might be described as chopped half waves—chopped at the isolated end of the winding. That is, the half wave is longer than the winding, and its extension beyond the winding is chopped off. In general, the higher the harmonic the greater percentage of the wave is chopped so that the oscillation approximates the quarter-wave mode. The lower the harmonic, the smaller is the fraction of the wave which is chopped, and thus the fundamental wave may be nearer to a half-wave length than to a quarter-wave length.

The leakage reactance of the transformer causes these modes of oscillation, a fact that cannot be ignored. If an ideal transformer were assumed with no reactance, magnetizing current, and so forth, no difference would be found between the oscillations of a winding as primary and those as secondary.

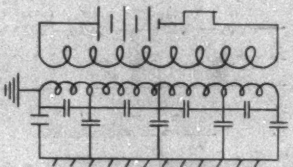

Figure 1. Essential elements of a transformer of the type here considered; considerable distributed capacitance in the secondary winding, negligible capacitance in the primary winding. Primary suddenly closed through a battery circuit

In the case of an oscillation that is intermediate between a quarter wave and a half wave, the maximum voltage to ground is in the interior, not at the terminal, of the winding. If the ratio of the maximum internal voltage (to ground) to the terminal voltage (to ground) be represented by μ_n for the nth harmonic and if the actual angular length of that oscillatory wave of the harmonic on the winding be represented by θ_n, then

$$\mu_n = 1/\sin\theta_n$$

This might be called the amplification factor of the harmonic voltage.

The angular length θ_n of the harmonic wave on the winding is found to be a certain function of the ratio of the nth-harmonic oscillatory leakage inductance L_n of the winding to the normal leakage inductance L_o of the winding, therefore if L_n/L_o equals 50, 20, 10, 5, 2, and 1, then μ_n equals respectively, 19, 6.6, 3.3, 2.2, 1.1, and 1. The conclusion follows that in order to keep μ_n low, the ratio L_n/L_o must be kept low.

The magnitudes of the various harmonic voltages for a rectangular wave are determined as follows:

1. Determine the angular length θ_n of each harmonic. This is a characteristic mode of the winding.
2. Determine the turn-ratio voltage distribution of the secondary winding, which will be a straight line stretching from zero at the grounded end to the maximum turn-ratio voltage at the other end of the winding. This sloping line is the normal (magnetic) initial voltage of the secondary winding. Because the presence of capacitance along the winding makes the actual initial voltage along the winding zero, the various harmonics at each point must initially add up (algebraically) to a value equal and opposite to the normal voltage at that point.
3. Choose as many points along the winding as the number of the harmonics desired to write enough simultaneous equations to determine the unknown magnitudes of the desired harmonics.

An example is shown in Figure 19.

The Circuit

Figure 1 illustrates the actual circuit. Its features are

1. One end of the high-voltage winding is grounded, the other isolated.
2. The high-voltage winding is associated with a network of coil-to-coil and coil-to-ground capacitances.
3. The voltage rating of the low-voltage winding is assumed so low that its associated capacitive kilovolt-amperes, varying as the square of the voltage, is negligible compared

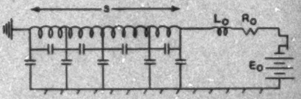

Figure 2. Tentative equivalent circuit of the transformer and of an analogous transmission line

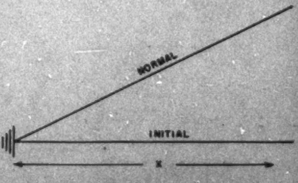

Figure 3. Initial and final distributions of rectangular voltage wave

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with that of the high-voltage winding. Hence no low-voltage capacitance network is shown.

4. The magnetizing current, whether primary or secondary, is ignored.

5. The resistances and leakage reactances are to be given quantitative consideration.

6. Excitation consists of a long-tail rectangular wave, like voltage from a battery, impressed across the low-voltage terminals.

7. The battery impedance is negligible.

Method of Analysis

The 2-winding system described can be replaced tentatively by an equivalent single-circuit network as shown in Figure 2. L_e is the leakage inductance between the two windings and R_e their effective resistance expressed in terms of the high-voltage circuit. Similarly, E_e is the battery voltage stepped up by the turn ratio. Although Figure 2 is not an exact equivalent of the transformer for these phenomena, it is very helpful in suggesting a simpler analogous problem in transmission lines to develop certain physical conceptions. After these conceptions have been developed the transformer equations will be based rigorously on Figure 1.

The problem is to formulate the transient phenomena of the high-voltage circuit following the closing of the battery switch, Figures 1 and 2.

Considering Figure 2, when the switch is closed the capacitance network evidently will behave as a short circuit momentarily and the entire voltage then will be consumed by the leakage inductance L_e so that the actual initial voltage distribution along the entire high-voltage winding will be zero, Figure 3. The normal voltage distribution, in the absence of the capacitance network, would have been along some slanting line like that shown in the same figure and will be the same for the secondary as for the primary winding, corrected for turn ratio. The normal voltage distribution of a winding frequently is spoken of also as the final distribution, and in the present case this may appear to be ultimately zero along all of the secondary winding. But as the transient is the result of the simultaneous magnetic and capacitive action and not the result of a prophetic vision of this winding into the distant future, a final voltage distribution can be a factor in calculation only in so far as it represents a component that was present at the beginning to co-operate with the other factors to control every detail of the transient phenomena. When so considered the desired reference axis for the oscillation will be recognized as the inductive-voltage distribution. The inductive voltage in the secondary winding does diminish with time but very slowly and for the present purpose may be considered as constant. This may be verified as follows.

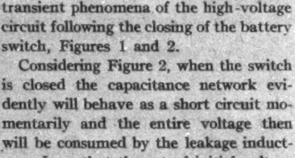
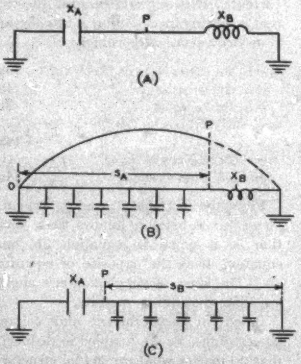


Figure 4. Standing wave in a transmission line which is grounded at both ends, as viewed from a point P on the line

The average value of a harmonic wave is 90 per cent of its rms value. Therefore if starting with zero flux in the core a battery voltage equal to the rated rms voltage is impressed across a 60-cycle transformer winding, it will be transformed magnetically to a secondary winding for at least a quarter of a cycle (4,000 microseconds) without drawing a magnetizing current sufficient to cause a noticeable voltage regulation through resistance drop. If a rectangular impulse voltage 20 times as large as the rated rms voltage of the winding is impressed under the same conditions, the inductive voltage transformation from the primary to the secondary winding will be as good as before for 200 microseconds. That is, it will be time enough for the maxima of the natural oscillations of many transformer windings to be exhibited fully.



Figures 5A, B, C. Three equivalent oscillatory networks of Figure 4 as viewed from the point P on the line

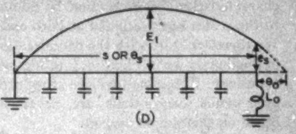


Figure 5D. Fundamental oscillatory wave length on a transmission line solidly grounded at one end and grounded through an inductance at the other end

The transition phenomena from initial to normal voltage distribution is then the crux of the problem. These will consist of a complex set of oscillations, but the question is what kind. So far as mathematics is concerned, the difference between the initial and the final curves of Figure 3 can be resolved into a set of harmonics in an infinite number of different ways, but only one of those sets would be acceptable, namely, that set of which each component constitutes a stable single-frequency mode of oscillation of the system. The modes of oscillation of the high-voltage as a primary winding as developed in reference 1 cannot be applied here unmodified for reasons as follows.

The fundamental mode of oscillation of the secondary as a secondary winding cannot be a half wave because

1. One end of the winding is open, Figure 1.

2. Such an oscillation in the secondary winding would not produce a voltage across the primary terminals and therefore it could not be produced in its turn by a voltage across the primary terminals. Also, the secondary oscillation cannot be a quarter wave because if the leakage reactance between the primary and secondary windings is zero or negligible, the oscillation would become short-circuited. If the leakage reactance is not negligible the oscillation will be short-circuited through that reactance and proportionately modified.

The first task evidently is to determine the normal modes of oscillation of the system in the presence of a leakage reactance L_e. In order to keep physics and mathematics together, and build the mathematics on the physics as far as possible, some pertinent basic physical conceptions first will be developed by considering the simpler problem of the analogous phenomena of a transmission line. These conceptions then will be developed further to adapt them to the transformer phenomena.

Analogous Problem in Transmission Lines

The system represented by Figure 2 is grounded at both ends and its oscillatory characteristics therefore must be related to those of a line grounded at both ends. We therefore start with the consideration of the line shown in Figure 4. It does not contain an L_o but that will be introduced after some conceptions are developed based on this simpler circuit.

Let the line of Figure 4 be assumed to oscillate at its fundamental mode—a half-cycle wave in space, not in time. The voltages to ground are zero at both ends, maximum in the middle. The conductor current is zero in the middle, maximum at the two ends, positive at one end and negative at the other. It should be borne in mind that we are dealing with a standing wave, not a traveling wave, which is alternating in time but stationary on the wire.

Let us make some observations and deductions at an intermediate point P on this line.

Looking from P into section A towards O, we find at P a voltage e_A and a current i_A.

$$e_A = E_1 \sin [2\pi(s_A/\lambda)] \cos [2\pi(c/\lambda)t] \quad \text{(1a)}$$
$$= E_1 \sin \theta_A \cos \omega t \quad \text{(1b)}$$
$$i_A = -(E_1/X_1) \cos [2\pi(s_A/\lambda)] \times \sin [2\pi(c/\lambda)t] \quad \text{(2a)}$$
$$= -(E_1/X_1) \cos \theta_A \sin \omega t \quad \text{(2b)}$$

where

E_1 = the maximum of the voltage wave
s_A = the length of section A in linear measure
θ_A = the length of section A in angular measure
λ = the wave length in linear measure
c = the velocity of wave propagation along the line in the same linear measure (per second) as used in measuring λ
X_1 = the surge impedance of the line

For ease of reference

$$\theta_A = 2\pi s_A/\lambda$$
$$\omega = 2\pi c/\lambda$$

The negative sign in equations 2a and 2b follows from the consideration that at a short distance from ground, θ_A small, the current must appear as lagging behind the voltage.

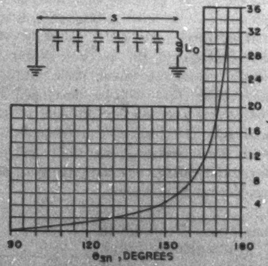
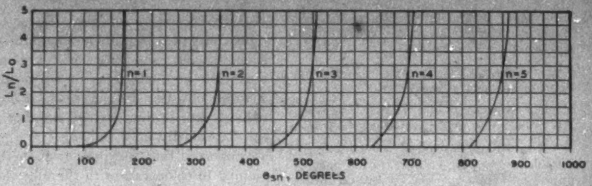


Figure 6A. Angular wave length of the fundamental oscillation of the line of Figure 5D



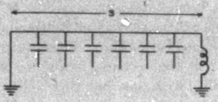


Figure 6B. Angular wave length of the various harmonic oscillations of the line of Figure 5D

Looking from P into section B toward the other end of the line, the same voltage will be found as before but an opposite current. It is opposite because the current is flowing from section A into section B or vice versa. Accordingly

$$e_B = E_1 \sin [2\pi(s_B/\lambda)] \cos [2\pi(c/\lambda)t] \quad \text{(3a)}$$
$$= E_1 \sin \theta_B \cos \omega t \quad \text{(3b)}$$
$$= E_1 \sin \theta_A \cos \omega t \quad \text{(3c)}$$
$$i_B = -(E_1/X_1) \cos [2\pi(s_B/\lambda)] \sin [2\pi(c/\lambda)t] \quad \text{(4a)}$$
$$= -(E_1/X_1) \cos \theta_B \sin \omega t \quad \text{(4b)}$$
$$= (E_1/X_1) \cos \theta_A \sin \omega t \quad \text{(4c)}$$

From the fact that θ_A and θ_B are supplementary angles it follows that equation 3c is equal to equation 3b, and equation 4c is the opposite of equation 4b. The sines of supplementary angles are alike, the cosines opposite.

Based on these current and voltage observations, we are justified in deducing that no matter what the actual structure of the two sections A and B of the circuit may be, viewed from the chosen point P, they offer the effective impedances X_A and X_B defined by

$$X_A = \frac{E_1 \sin \theta_A}{I_1 \cos \theta_A} \quad \text{(5a)}$$
$$= X_1 \tan \theta_A \quad \text{(5b)}$$
$$= -X_1 \tan \theta_B \quad \text{(5c)}$$
$$X_B = \frac{E_1 \sin \theta_B}{I_1 \cos \theta_B} \quad \text{(6a)}$$
$$= X_1 \tan \theta_B \quad \text{(6b)}$$
$$= -X_1 \tan \theta_A \quad \text{(6c)}$$

X_1, equal to E_1/I_1, will be recognized as the surge impedance of the line to the fundamental wave of the line. Although the surge impedance of a transmission line is the same to all waves, for the sake of generality (to be utilized later in transformer windings) the subscript will be retained.

The signs of these equations are so chosen that positive X means a lagging reactance, negative leading.

If θ_A is more than 90 degrees as shown in Figure 4, X_A, the reactance of section A effective at the point P, will be negative, that is, equivalent to a capacitive reactance. θ_B will be less than 90 degrees and therefore X_B will be positive, equivalent to an inductive reactance.

Thus considered from the standpoint of our chosen point of observation P, the oscillating system can be represented by any one of three equivalent networks.

1. Figure 5A consisting of a massed capacitive reactance X_A and a massed inductive reactance X_B.
2. Figure 5B consisting of a distributed line of length s_A at the left terminating through a massed inductive reactance X_B at the right.
3. Figure 5C consisting of a massed capacitive reactance X_A at the left and a line of length s_B on the right.

Conversely, if we have an actual circuit as illustrated in Figures 5A, B, or C, it can be represented by a transmission line of length $\lambda/2$, Figure 4, with no massed reactance anywhere.

Fundamental Mode of Oscillation of a Line Terminating in a Massed Inductance

Several important general conclusions and novel concepts follow from the foregoing text.

(*a*). The fundamental mode of oscillation of a line of linear length s, Figure 5D, terminating through a massed inductance L_o and grounded at both ends is not a half wave but a fractional or stub-ended wave of angular value θ_s given by

$$1/\tan \theta_s = -X_1/\omega L_o \quad \text{(7a)*}$$
$$\theta_s/\tan \theta_s = -L_1/L_o \quad \text{(7b)*}$$

L_1/L_o is the ratio of the distributed inductance L_1 of the line to the massed inductance L_o at the end. The values of θ_s corresponding to given values of L_1/L_o can be obtained from Figure 6A.

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*Note that X_1, is the surge impedance of the line, not its inductive reactance. Therefore X_1, is not equal to ωL_1 but to $\omega L_1/\theta_s$, as defined by equation 7e.

Equations 7a and 7b are derived from equation 6c by substituting into the latter the following relationships.

$$X_B=\omega L_o;\ \omega=2\pi c/\lambda;\ \lambda=2\pi s/\theta_s$$

Therefore

$$\omega=c\theta_s/s$$

Let L and C be the line constants per unit length, so that L_1 equals sL. Then

$$c=1/\sqrt{LC}$$

Therefore

$$C=1/Lc^2 \tag{7c}$$

$$\begin{aligned}X_1&=\sqrt{L/C}\\&=Lc\end{aligned} \tag{7d}$$

Hence

$$\begin{aligned}X_1/\omega&=Lc/(c\theta_s/s)\\&=L_1/\theta_s\end{aligned} \tag{7e}$$

Substituting equation 7e into equation 6c and remembering that X_B is also equal to ωL_o, we obtain

$$\theta_s/\tan\theta_s=-L_1/L_o \tag{7b}$$

Just as θ_s is the portion of the wave (in angular measure) subtended by the line, $\theta_o(\theta_o=\pi-\theta_s)$ is the portion of the wave subtended by the massed inductance L_o. Substitution of $\theta_s=\pi-\theta_o$ into equation 7b yields

$$(\pi-\theta_o)/\tan\theta_o=+L_1/L_o \tag{8}$$

With increasing values of L_1/L_o, θ_o becomes smaller approximating the value

$$\theta_o\cong\frac{1}{L_1/L_o} \tag{8a}$$

As equations 7b and 8 are solved for θ_s and θ_o conveniently with the help of the curves of Figure 6A, equation 8a need not be resorted to except for a rough estimate.

It can be seen from either the equations or the curves that no matter how large L_o may be, its angle θ_o will be less than 90 degrees and θ_s between 90 and 180 degrees.

As a matter of theoretical interest, if the line terminates through a massed capacitance instead of inductance, these relations between the two θ's will be interchanged. θ_s will be between 0 and 90 degrees, θ_o between 90 and 180 degrees.

(*b*). A line of length s terminating through a massed inductance and grounded at both ends oscillates as if it were part of a line of a greater length s' supporting a full half wave.

$$\begin{aligned}s'&=s(\pi/\theta_s)\\\lambda&=2s'\end{aligned} \tag{9a}$$

(*c*). The frequency of oscillation of such a standing wave is given by

$$\begin{aligned}\omega&=2\pi c/\lambda\\&=c\theta_s/s\end{aligned} \tag{10}$$

This means that series massed inductance lowers the frequency of oscillation of a line in the ratio of π to θ_s.

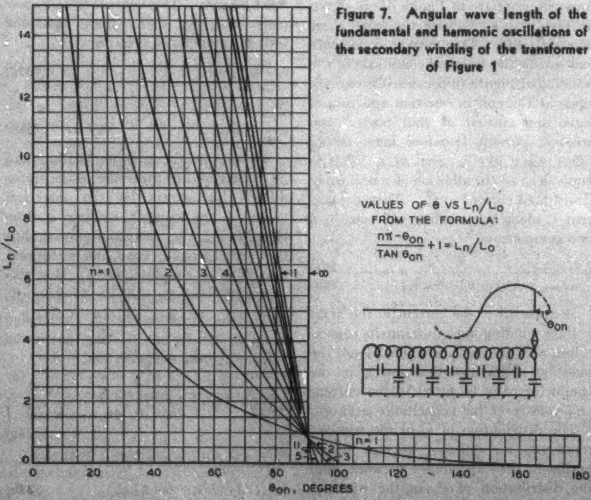


Figure 7. Angular wave length of the fundamental and harmonic oscillations of the secondary winding of the transformer of Figure 1

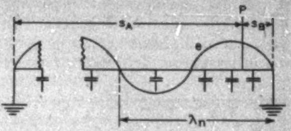


Figure 8. Harmonic wave on a transmission line grounded at both ends

(*d*). If the circuit conditions are such as to fix the line terminal voltage e_s Figure 5D and to compel E_1 to adjust itself to it, then there will be a voltage amplification along the line in the ratio of E_1 to e_s given by the equation

$$\mu_1=E_1/e_s=1/\sin\theta_s=1/\sin\theta_o \tag{11}$$

Since in equation 8 the larger the ratio L_1/L_o the smaller is θ_o and for small values of θ, $\sin\theta=\theta=\tan\theta$, the substitution of this into equations 8 and 11 yields the approximation

$$\mu_1\approx L_1/(L_o\pi) \tag{11a}$$

That is, the higher values of harmonic amplification are of the order of one-third of L_1/L_o.

HIGHER HARMONIC MODES

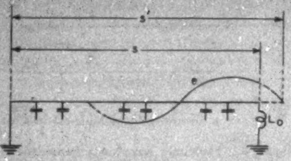
In Figure 8 the transmission line of Figure 4 is shown as oscillating under a standing wave of the nth space-harmonic of wave length λ_n. Observations made as before will lend to the same general equations as before for current and voltage and therefore for X_A, X_B, θ_s (equation 7b), θ_o (equation 8), and μ (equation 11) by substituting the appropriate value of λ for the desired harmonic and by multiplying π by n in equation 8.

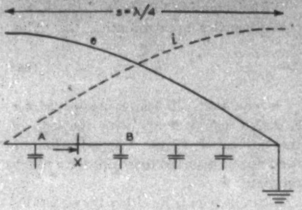
Of major interest are θ_o and μ for the line of Figure 5D for the nth harmonic.

Equation 7b for θ_s does not seem to have provision for different harmonics although it does inasmuch as it has an infinite number of solutions, one for each harmonic. Figure 6A shows only a fragment of the complete plot—just enough to cover the fundamental mode. Figure 6B shows a more extended one with the curves of θ_s up to the fifth harmonic; and Figure 7 is in terms of θ_o up to the eleventh harmonic. Thus it is seen that equation 7b (or equation 8 modified) defines all of the infinite number of different possible normal modes of oscillation of a line terminating through a massed inductance and grounded at both ends.

GENERAL FORMULAS

We now put the more important formulas in a form applicable to all harmonics of the transmission line, including the





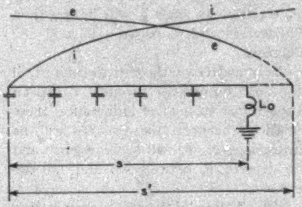


Figure 9 (above). Harmonic wave on a transmission line solidly grounded at one end and grounded through an inductance at the other end

Figure 10 (top right). Fundamental wave on a line open at one end, grounded at the other end

Figure 11 (bottom right). Fundamental wave on the line open at one end, grounded through an inductance at the other end

fundamental (half wave) for which $n=1$.

$$\theta_{sn}/\tan\theta_{sn}=-L_n/L_e \tag{12}$$

and

$$\theta_{on}=n\pi-\theta_{sn} \tag{13}$$

θ_{sn} can be obtained more conveniently from Figure 6B, or

$$\frac{n\pi-\theta_{on}}{\tan\theta_{on}}=+L_n/L_e \tag{14}$$

and

$$\theta_{sn}=n\pi-\theta_{on} \tag{15}$$

Also

$$\begin{aligned}\mu_n&=1/\sin\theta_{sn} \qquad (16)\\&=1/\sin\theta_{on}\\ \omega_n&=c\theta_{sn}/s\\&=\theta_{sn}/(s\sqrt{LC}) \qquad (17)\\&=c(n\pi-\theta_{on})/s\\ s'_n&=s(n\pi/\theta_{sn}) \qquad (18)\\ \lambda_n&=2s'_n/n\\&=2\pi s/\theta_{sn}\end{aligned}$$

The fact that the equations for θ_{on} and θ_{sn} are transcendental forbids any expectation that s'_n, that is, the equivalent length of the system will be the same for all harmonics.

Open Line

The stable modes of oscillation of an open line like those shown in Figures 10 and 11 are analogous to those of the isolated high-voltage winding and are analyzed in Appendix II. Equations 20–27 also will be found there.

Other Combinations

A line may have stable modes of oscillation with massed reactances introduced at intermediate points as in Figures 14 and 15. These are discussed in Appendix III.

Applications to the Transformer

Constraints of the Transformer

It will be shown now how a fractional standing wave, like that of Figure 5D for instance, is also the normal mode of oscillation of the secondary winding of a transformer satisfying its current and voltage constraints.

Current Constraints of the Secondary. The voltage wave of Figure 5D is reproduced in Figure 16 as e_x. The current wave appropriate to Figure 5D is shown in Figure 16 as i'_x with the value i'_s at the stub end. This current wave does not satisfy the terminal conditions of the winding of Figure 16 because it is entirely open at the end in question and cannot have any current at that point. The winding current therefore must be an offset wave like i_x, zero at s. This is equivalent to the addition of a uniformly distributed current $-i'_s$. Then the current i_x along the winding will consist of two components.

$$\begin{aligned}i_x&=i'_x+i''_x \qquad (29a)\\&=I_{\max}\{\cos[(x/s)\theta_s]-\cos\theta_s\} \qquad (29b)\end{aligned}$$

Because all of the current in the high-voltage winding must flow into its capacitance network as a distributed load, the question arises how this new i_x in the conductor is to be fitted to the current and voltage of the capacitance network.

The distribution of i_x in the winding represents no different current distribution in the capacitance network than the distribution of i'_x in the winding. The former means that the current of the right-hand portion of the capacitance network, instead of flowing through the winding toward the right (toward the open end) to return, flows toward the left (the closed end) to return to the ground.

The component i'_x produces the voltage e_x while i''_x contributes to the leakage reactance voltage across the primary or low voltage.

Primary (Low-Voltage) Current. A complete half wave of current in the high-voltage winding, like that shown in Figure 4, would call for no current i_p in the low-voltage primary winding, but one similar to either the unsymmetrical i'_x (with unequal half cycles) or the offset i_x would do so, by the general formula,

$$i_p=-(1/s)\int_0^s i_x dx \tag{30a}$$

As all turns are in series in the primary winding and there are no shunt paths, this current i_p will flow in all of its turns uniformly, not as a function of x.

Substituting equation 29b into equation 30a and integrating,

$$i_p=-I_{\max}\left[\frac{\sin\theta_s}{\theta_s}-\cos\theta_s\right] \tag{30b}$$

and

$$\begin{aligned}i_p/i_s&=-i_p/I_{\max}\cos\theta_s\\&=-\left(\frac{\tan\theta_s}{\theta_s}-1\right) \qquad (31)\end{aligned}$$

As the low-voltage terminals are connected to a zero-impedance source, the total oscillatory voltage at its terminals must be zero. This consists of two components: e_s and $i_p\omega L_e$ where L_e is the leakage inductance. Hence

$$e_s+i_p\omega L_e=0 \tag{32a}$$

or

$$e_s=-i_p\omega L_e \tag{32b}$$

We see that the same applications can be made here as in the case of the transmission line except for the numerical value of the current causing a drop through L_e. In the transmission line the current was $I_{\max}\cos\theta_s$ and in this case it is given by equation 30b. How this affects θ_e and θ_s will be determined.

Equations for θ_e and θ_s. Developing equation 32b by substituting into it the value of i_p from equation 30b,

$$e_s=E_1\sin\theta_s=(E_1/X_1)\left[\frac{\sin\theta_s}{\theta_s}-\cos\theta_s\right]\omega L_e \tag{33a}$$

or dividing through by $E\sin\theta_s$

$$1=(1/X_1)[1/\theta_s-\cot\theta_s]\omega L_e \tag{33b}$$

$$X_1/\omega L_o = (1/\theta_s) - (1/\tan\theta_s) \qquad (33c)$$

Equation 7e stated

$$X_1/\omega = L_1/\theta_s$$

which when substituted into equation 33c and interpreted for the *n*th harmonic yields

$$(\theta_{sn}/\tan\theta_{sn}) - 1 = -L_n/L_o \qquad (34)$$

and

$$\frac{n\pi - \theta_{on}}{\tan\theta_{on}} + 1 = (L_n/L_o) \qquad (34a)$$

Given L_n/L_o, θ_{on} can be obtained conveniently from Figure 7.

L_n is involved also in the formula for the frequency of the oscillation of the harmonic and will be formulated under that topic. Meanwhile for the purposes of equations 34 and 34a, the following thought may be found helpful.

In these equations the ratio, not the absolute values, of L_n and L_o is needed. Therefore factors like $(4\pi/10)$ that enter into all inductance formulas can be cancelled out from both numerator and denominator leaving finally

$$L_n/L_o = 1.25\frac{A_n}{A_o} \times \frac{h_o}{h_n}$$

where

A = the effective cross section of the leakage path
h = the effective length of the leakage path
1.25 = a factor derived in the following text

The impedance condition to which A_n and h_n apply are illustrated in Figure 21 for the first three harmonics with current-flow diagrams familiar in transformer design practice.

Comments on Figure 7. Figure 7 shows that, with varying values of L_n/L_o, θ_{on} varies over a wide range for the fundamental, and narrower ranges with the increasing order of the harmonics. As L_n/L_o itself gets smaller with the increasing order of the harmonics for a given winding, it follows that the reduction in range is even more pronounced than it appears at first sight. For instance, to a rough approximation (Appendix V) L_n may be assumed to vary inversely as the square of the order of the harmonic, and on that basis the exhibit in Table I for two values of L_1/L_o may be enlightening.

It is seen that the oscillation approaches the quarter-wave type for all harmonics above the fundamental in case 2 and above the second in case 1. It can be less than $\pi/2$.

Aside from the visual picture this gives of the character of the various harmonics, it also affords an opportunity for simplification of the harmonic analysis of complex waves by approximation.

GENERAL EQUATIONS OF THE HIGH-VOLTAGE (SECONDARY) WINDING

Based on the foregoing analyses the equation of the voltage to ground along the secondary winding, at a distance x from neutral, with a rectangular voltage impressed across the primary, and with allowance also for damping, will be of the form

$$e_x = E_o(x/s) + \sum_1^n E_n \epsilon^{-k_n t} \sin[(x/s)\theta_{sn}] \times \cos\omega_n t \qquad (35)$$

This involves the unknowns E_n, k_n, θ_{sn}, and ω_n. Of these θ_{sn} already has been defined and is obtainable from Figure 7 in terms of θ_{on}. As k_n, the damping exponent, is influenced strongly by stray inductive and dielectric losses difficult to formulate, it is better to estimate it from experience. That leaves two unknowns to formulate—E_n and ω_n.

Amplitude of Harmonics, E_n

Given the initial actual and the normal voltage distributions Fig. 3, how are the amplitudes of the fundamental and the higher harmonics to be determined?

The well-known Fourier analysis does not apply here because, if the integration is done for the actual winding length s, the waves are incomplete. If the integration is done for s', that is, the equivalent length, then s' is not the same for each harmonic.

Still, the harmonics can be obtained by one of several different methods.

Method 1. Calculate θ_{Lo} (or θ_s) for as many harmonics as desired and plot the waves, each with 100-per-cent maximum amplitude, together with the final voltage distribution curve. If this is done on co-ordinate paper, then at any point along the winding it is seen at a glance that a certain percentage of the fundamental harmonic plus certain other percentages of the various harmonics at the first instant must be equal and opposite to the local value of the final voltage distribution. Therefore with the help of the plot equations may be written for as many points of the winding as the number of the simultaneous equations needed, that is, the number of the harmonics.

Table I

Harmonic	Case 1: $L_1/L_o = 9$		Case 2: $L_1/L_o = 5$	
	L_n/L_o	θ_{on}, Degrees	L_n/L_o	θ_{on}, Degrees
Fundamental	9	20	5	32
Second	2.25	76	1.25	88
Third	1.0	90	0.55	94
Fourth	0.56	95	0.31	93
Fifth	0.36	93	0.25	92
Infinite	0.00	90	0.00	90

Method 2. A set of simultaneous equations can be written without the initial plot of method 1 as follows.

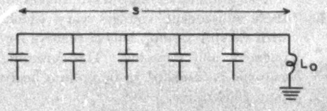
Let it be assumed that only three harmonics are required. At $x = s$, the transient component of $e_x = -E_o$. Hence

$$E_1\sin\theta_{o1} + E_2\sin\theta_{o2} + E_3\sin\theta_{o3} = -E_o \qquad (39)$$

In all cases at $x = s'_2/2$ the second harmonic is zero, and the transient component of voltage is $-(s'_2/2s)E_o$. Hence

$$E_1\sin[(s'_2/2s)\theta_{s1}] + E_3\sin[(s'_2/2s)\theta_{s3}] = -(s'_2/2s)E_o \qquad (40)$$

Finally, at $x = s'_3/3$, the third harmonic is zero, and the transient component of



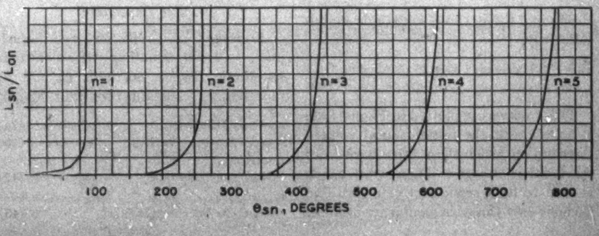


Figure 12. Angular wave length of the various harmonic oscillations of the line open at one end and grounded through an inductance at the other end

the voltage is $-(s'_3/3s)E_o$. Hence

$$E_1 \sin [(s'_3/3s)\theta_{s1}] + E_2 \sin [(s'_3/3s)\theta_{s2}] = -(s'_3/3s)E_o \quad (41)$$

Thus we have three equations to determine the three unknowns E_1, E_2, and E_3.

When few harmonics are considered, for example three, both methods are approximate and the answers depend on the choice of points used for setting up the equations.

Example — Calculation of Harmonics. For simplicity only three harmonics will be considered.

Let

$L_1/L_o = 5$
$L_2/L_o = 2.8$
$L_3/L_o = 1$

Then

$\theta_{o1} = 30°$
$\theta_{o2} = 70°$
$\theta_{o3} = 90°$
$\theta_{s1} = 150°$
$\theta_{s2} = 290°$
$\theta_{s3} = 450°$
$\sin \theta_{s1} = 0.5$
$\sin \theta_{s2} = -0.94$
$\sin \theta_{s3} = 1$

The plotting of the harmonics by method 1 is facilitated by laying out the scale of angles for each harmonic as shown in Figure 19A. The bottom scale is for the fundamental harmonic, the middle scale for the second harmonic, and the top scale for the third harmonic. The waves with 100 per cent amplitude are not reproduced there in order to avoid confusion with the actual curves of the harmonics.

At the open end

$$E_1 \sin 150° \text{ (or } 30°) + E_2 \sin 290° \text{ (or } 70°) + E_3 \sin 450° \text{ (or } 90°) = -100 \text{ per cent}$$

That is,

$$0.50E_1 + 0.94E_2 + E_3 = -100 \text{ per cent} \quad (39a)$$

At 60, the fundamental degrees, the third harmonic is zero. Hence

$$0.86E_1 - 0.9E_2 = -40 \text{ per cent} \quad (39b)$$

Finally, at 90, the fundamental degrees,

$$-E_1 + 0.1E_2 + E_3 = -60 \text{ per cent} \quad (39c)$$

Solving these three equations simultaneously,

$$E_1 = -85 \text{ per cent}; \; E_2 = 38 \text{ per cent}; \; E_3 = -22 \text{ per cent} \quad (40)$$

These waves are plotted in Figure 19A.

As a check the initial actual voltage distribution which these three harmonics represent is shown in Figure 19B which is seen to be a crude sort of a zero line because only three harmonics are utilized.

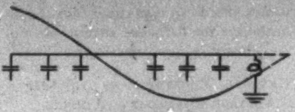


Figure 13. Higher harmonic wave on the line open at one end, grounded through an inductance at the other end

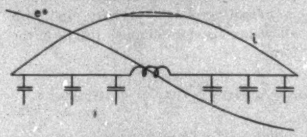


Figure 14. Oscillation of a line having a massed series inductance in the middle

Example—Voltage Distribution a Half Cycle Later. To determine the resultant voltage distribution a half cycle (of the fundamental) later, let it be assumed first that $\omega_2 = 3\omega_1$, that the third harmonic is almost damped out, and the fundamental and second harmonic are substantially undamped. The result is shown in curve *A* of Figure 19C. The maximum voltage to ground is 188 per cent of the impressed (turn-ratio) voltage. The maximum gradient is near the middle and is about 3.2 times that corresponding to uniform distribution.

As an alternative and more realistic assumption, let it be assumed that the third harmonic will be damped out 100 per cent, the second harmonic 75 per cent, and the fundamental 30 per cent. The resulting distribution is shown by curve *B*, Figure 19C. The maximum voltage is now 142 per cent instead of 188 per cent; and the maximum gradient is 2.2 instead of 3.2 times that corresponding to uniform distribution of the turn-ratio impulse voltage.

The amplitudes and wave shapes of these subsequent voltages vary widely with varying assumptions of relative frequencies and damping. The foregoing example is intended to be nothing more than illustrative.

Frequency of Oscillation, Inductance, and Surge Impedance

Frequency of Oscillation and Inductance. In a standing wave of unit amplitude, Figure 20, with C_s and C_g representing constants per unit length,

$$e_g = \sin [(2\pi/\lambda)x] \quad (41a)$$
$$e_s = (2\pi/\lambda) \cos [(2\pi/\lambda)x] \quad (41b)$$
$$i_g = \omega C_g e_g = \omega C_g \sin [(2\pi/\lambda)x] \quad (42)$$
$$i_s = \omega C_s e_s = \omega C_s (2\pi/\lambda) \cos [(2\pi/\lambda)x] \quad (43)$$
$$i_L = i_s + \int_x^{\lambda/4} i_g dx \quad (44a)$$
$$= \omega(C_s 2\pi/\lambda + C_g \lambda/2\pi) \cos [(2\pi/\lambda)x] \quad (44b)$$

It may be recognized in equation 44b that $C_s(2\pi/\lambda)$ represents the series capacitance per radian and may be symbolized by C'_s. Similarly, $C_g(\lambda/2\pi)$ is ground capacitance per radian and may be symbolized by C'_g. Equation 44b then may be written as

$$i_L = \omega(C'_s + C'_g) \cos [(2\pi/\lambda)x] \quad (44c)$$

It may be interesting to observe that when the series and ground capacitances are entered per radian they enter additively to each other with equal weight, as equivalent to each other, or interchangeable, so to say. We therefore might represent $C'_s + C'_g$ by the single symbol C' as the total effective capacitance per radian.

The physical meaning of this is that a C_s distribution of one microfarad per radian constitutes identically the same burden on the winding as a C_g distribution of one microfarad per radian.

Now we must determine what value of ω (if any) will enable the i_L of equation 44c to produce the e_s of equation 41b through whatever inductive effect the winding may offer to it.

Let

N = the number of turns per unit length
h = the effective length of the harmonic-flux path

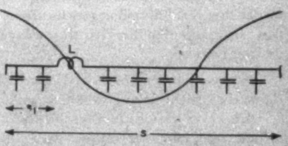


Figure 15. Oscillation of a line having a massed series inductance at a voltage node other than midpoint

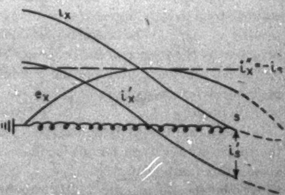


Figure 16. Components of current wave to satisfy the primary and secondary terminal conditions of Figure 1

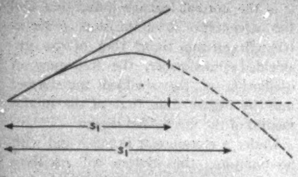


Figure 17. Actual fractional wave and corresponding complete half wave

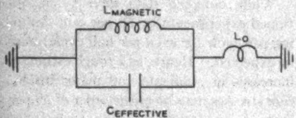


Figure 18. Equivalent inductance and capacitance network of shielded transformer

p = the mean perimeter of the winding

Then

$$B=(4\pi/10)(N/h)\int_0^x i_L dx \qquad (45)$$

$$d\phi=1.26(Np/h)\int_0^x i_L dx \qquad (46)$$

$$\frac{de_s}{dx}=1.26\times10^{-8}(N^2p/h)\omega\int_0^x i_L dx \qquad (47)$$

$$e_s=1.26\times10^{-8}(N^2p/h)\omega\int_x^{\lambda/4}\int_0^x i_L dx^2 \qquad (48a)$$

Substituting the value of i_L from equation 44c,

$$e_s=1.26\times10^{-8}(N^2p/h)\omega^2(C'_s+C'_g)\times \int_x^{\lambda/4}\int_0^x \cos[(2\pi/\lambda)x]dx^2 \qquad (48b)$$

$$=1.26\times10^{-8}(N^2p/h)(\lambda/2\pi)^2\omega^2(C'_s+C'_g) \cos[(2\pi/\lambda)x] \qquad (48c)$$

If e_s is multiplied by $\lambda/2\pi$, it becomes volts per radian. Doing this to both members of equation 48c,

$$(\lambda/2\pi)e_s=1.26\times10^{-8}(N^2p/h)\times (\lambda/2\pi)^3\omega^2(C'_s+C'_g)\cos[(2\pi/\lambda)x] \qquad (49a)$$

$$=\omega L'\omega(C'_s+C'_g)\cos[(2\pi/\lambda)x] \qquad (49b)$$

$$=\omega L' i_L \qquad (49c)$$

in which

$$L'=1.26\times10^{-8}(N^2p/h)(\lambda/2\pi)^3 \qquad (49d)$$

From equation 49c it is clear that L' is the effective inductance of the winding per radian under a sinusoidal current wave.

To solve for ω we equate the e_s in equation 41b to the e_s in equation 49b. Cancelling similar factors from both sides leaves

$$1=\omega^2L'(C'_s+C'_g) \qquad (50)$$

Hence

$$\omega=\frac{1}{\sqrt{L'(C'_s+C'_g)}} \qquad (51a)$$

$$=\frac{1}{\sqrt{L'C'}} \qquad (51b)$$

where in per-radian quantities

$$C'=C'_s+C'_g \qquad (51c)$$

Because the values of both L' and C' vary with the harmonic, it is better to write the constant with the subscript n to make this thought more explicit so that equation 51b would be written as

$$\omega_n=\frac{1}{\sqrt{L'_nC'_n}} \qquad (51d)$$

Because L'_n is the inductance per radian and L_n in previous equations was the total inductance of the winding, it follows that

$$L_n=\theta_{sn}L'_n \qquad (51e)$$

where θ_{sn} is in radians.

It may be recognized that equations 51b and 51d also are applicable to transmission lines. By expressing the constants of the system per radian, we have secured a generality of formula applicable to any harmonic whether on a transmission line or on a transformer winding. This also is true of surge impedance.

Inasmuch as it is not practical to secure a sinusoidal distribution of current for the experimental determination of the inductance at low frequencies, it may be well to determine the relationship of L' to the nearest quantity that can be determined by test and with the conception of which we may be familiar.

The nearest familiar inductance concept to L' is the leakage inductance L'' corresponding to an ampere-turn distribution that has the same wave length but is triangular instead of sinusoidal, with equal current in each turn. A half group with triangular magnetomotive-force distribution corresponds to (though does not equal) the sinusoidal half wave. We assume that the current is one ampere in each turn in the triangular magnetomotive-force case and one ampere maximum in the sinusoidal case.

The leakage inductance L'' of such a half group of length $\lambda/2$ with triangular ampere-turn distribution is

$$L''=1.26\times10^{-8}(N^2p/h)\lambda^3/96 \qquad (52)$$

L'' can be either calculated or measured. Calculation may sometimes involve the uncertainty of the effective value of h and direct measurement then is desirable.

As the half group corresponds to π ra-

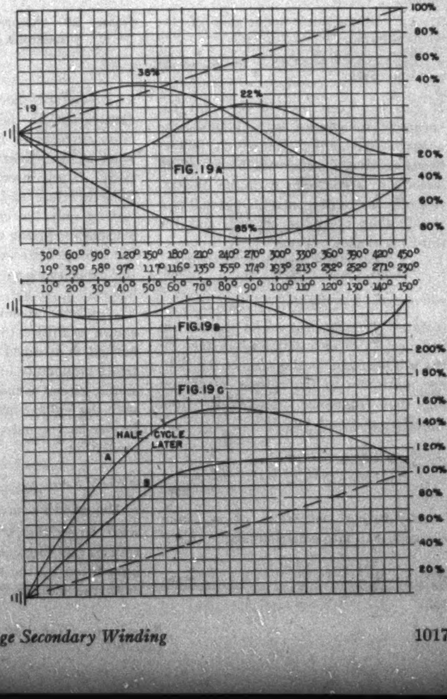


Figure 19

A. Angular wave length of three harmonics of a transformer

B. The 3-harmonic initial resultant voltage distribution

C. Voltage distribution a half cycle after start dians, the equivalent value of the inductance per radian will be the L'' of equation 52 divided by π.

$$L''_{\text{per radian}} = L''/\pi \tag{53a}$$

$$= 1.26\times10^{-8}(N^2p/h)\lambda^3/96\pi \tag{53b}$$

Comparing equations 53 and 50, it may be seen that

$$L' = (12/\pi^2)L''_{\text{per radian}} = 1.2515L''_{\text{per radian}} \tag{54}$$

This is an interesting result. The effective inductance per radian for a sinusoidal distribution of magnetomotive force is $12/\pi^2$ or 1.2515 times that for a triangular distribution of magnetomotive force.

Based on equations 51e and 54, L_n for the whole winding is given by

$$L_n = 1.25L''_{\text{per radian}}\times\theta_{sn} \tag{55a}$$

$$= 1.25L''\theta_{sn}/\pi \tag{55b}$$

When θ_{sn} is almost a whole multiple of π, there are n half groups and

$$L_n = 1.25nL'' \tag{55c}$$

When making measurement it is desirable to do so for complete waves and then interpolate for fractional waves.

Measuring the Inductance. Three inductances have been discussed.

1. L'_n, the harmonic inductance per radian.
2. L_n, the harmonic inductance of the whole winding.
3. L'', the inductance of a half-group for uniform current.

It also has been pointed out that the harmonic inductance is 1.25 times the uniform-current inductance. It follows therefore that $L_n/1.25$ can be measured directly in accordance with the illustrative impedance diagrams of Figure 21 for the fundamental, the second, and the third harmonics. Then

$$L'_n = L_n/\theta_{sn} \tag{55d}$$

where θ_{sn} is in radians.

Surge Impedance. The surge impedance Z_n of an oscillatory system for the nth harmonic is the ratio of the voltage maximum to the current maximum for that harmonic. Therefore, from equations 41a and 44c,

$$Z_{n\ \text{surge}} = \frac{e_{g\ \max}}{i_{L\ \max}} = \frac{1}{\omega_n(C'_s+C'_g)_n} = \frac{1}{\omega_nC'_n} \tag{56a}$$

Substituting the value of ω_n from equation 51d into equation 56a,

$$Z_{n\ \text{surge}} = \sqrt{\frac{L'_n}{C'_n}} \tag{56b}$$

which also applies for transmission lines and for all harmonics. This generality also is secured by expressing the constants per radian. In the transmission line the values of L' and C' vary similarly with wave length and therefore their ratio, the surge impedance, is constant for all harmonics. In the transformer winding the values of L' and C' vary as different functions of the wave length and accordingly the surge impedance varies with the harmonic. It is the more creditable for equation 56b therefore that it fits both cases.

Oscillations of a Shielded Secondary

Because a shielded secondary maintains the same voltage distribution along its capacitance network as along its inductive circuit, there is no tendency for cross currents to flow from one to the other at their intermediate cross connections. These intermediate cross connections therefore may be opened (on paper) leaving only the end connections intact without modifying the phenomena. With the intermediate cross connections open, the capacitance network becomes equivalent to a single massed capacitance C_{eff} and the system equivalent to that shown in Figure 18. If the magnetizing current is assumed zero and the magnetizing inductance infinite, then we have a simple circuit consisting of C_{eff} in series with the leakage inductance of the transformer having only one possible mode of oscillation

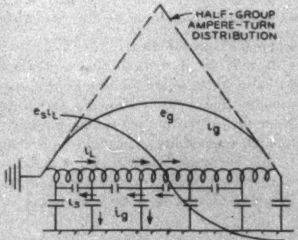


Figure 20. Ampere turn distributions—sinusoidal and triangular

$$\omega = 1/\sqrt{L_sC_{eff}} \tag{57}$$

The effective capacitance will be somewhat larger than $\sqrt{C_gC_s}$ of the unshielded transformer because of the capacitance introduced by the shielding.

As the inductance in the formula for ω is the normal leakage inductance and the capacitance is of the same order as (though greater than) that of the unshielded transformer, the frequency of oscillation of the shielded transformer may be higher than that of the fundamental of the unshielded design.

Under a rectangular impulse and with no damping, this system will oscillate between zero voltage and double the impulse voltage, both to ground and across the coils, maintaining uniform distribution all the time.

While damping data have to be obtained experimentally, 50 per cent damping per cycle (or even per half cycle) may be expected. Hence, as a rough estimate, increases in gradients and major insulation stresses may be of the order of 50 per cent.

Inasmuch as all internal stresses of a shielded design are proportional to its terminal voltage, whenever it is assured that the terminal voltages will be held down to a safe value, no greater concern need be felt for the simple shielded winding when acting as the secondary than when acting as the primary winding.

Appendix I. Velocity of Propagation in Transformer Windings

The concept of velocity of propagation is ill-fitting to the concept of a standing wave—bad enough in the case of a transmission line, worse in that of the transformer windings. Therefore any symbol that may be taken for velocity of propagation will be interpreted in terms of standing waves.

In some of the early equations of a transmission line, a velocity of propagation c was utilized because originally (in classical analyses) the standing wave is derived as the resultant of two waves traveling in opposite directions. But after the concept of a standing wave has been established, c may as well be given a new interpretation better fitting the standing wave concept. As c appears in our formulas for ω, and $\omega = 2\pi f = 2\pi c/\lambda$, it follows that $c = f\lambda$; and for $f = 1$, $c = \lambda$. That is, c is the wave length of a standing wave oscillating at one cycle per second.

If the frequency of oscillation has constant proportionality to the reciprocal of the wave length, as in a transmission line, the foregoing formula gives the same value of c reckoned from any harmonic $c = f_n\lambda_n$. Since the ratio of f to $1/\lambda$ is not constant for all wave lengths in the case of windings, c will have a different value c_n as reckoned from different harmonics, $c_n = f_n\lambda_n$.

The meaning of c_n in windings then is that wave length which a standing wave would have to have in order to oscillate at one cycle per second if the winding had linear characteristics proportional to those of the nth harmonic. Thus the 1-cycle-per-second wave length is different as reckoned from different harmonics.

Appendix II. Transmission Line Completely Open at One End

This is a further discussion of "Open Line" of the text. The circuit is illustrated in Figure 10 and is analogous to the isolated high-voltage winding.

Fundamental Mode of Oscillation

The standing wave of the longest possible wave length on such a line is a quarter of a wave length ($\lambda=4s$), and the voltage and current equations referred to the open end as origin are

$$\left.\begin{aligned} e_x &= E_1 \cos\,[2\pi(x/\lambda)]\cos\,[2\pi(c/\lambda)t] \\ &= E_1 \cos\theta_x \cos\omega_1 t \end{aligned}\right\} \quad (20)$$

$$\left.\begin{aligned} i_x &= -(E_1/X_1)\sin\,[2\pi(x/\lambda)] \\ &\qquad \sin\,[2\pi(c/\lambda)t] \\ &= -(E_1/X_1)\sin\theta_x \sin\omega_1 t \end{aligned}\right\} \quad (21)$$

The algebraic sign is so assumed as to make the current lead the voltage.

At any point x along the line, the effective reactances of the two sections of the line will be

$$X_A = e_x/i_x = -X_1\cot\theta_x \quad (22)$$

$$X_B = -X_A = X_1\cot\theta_x \quad (23a)$$

$$= -X_1\tan\left(\frac{\pi}{2}+\theta_x\right) \quad (23b)$$

As $\theta_x<90$ degrees and all trigonometric functions are positive in the first quadrant, X_A will be negative or capacitive and X_B positive or inductive.

A massed inductance connected to the open end (but without closing the circuit to ground) can have no current and no voltage drop and therefore no effect whatever on the oscillation of the line. However, if inserted at the grounded end, Figure 11, it becomes equivalent to an extension of the line from s, the actual length, to s', an equivalent length, derived as follows.

If the terminating reactance is capacitive, the circuit cannot oscillate to a quarter wave length of any modified dimensions because two capacitive reactances cannot oscillate—cannot resonate—with each other. Oscillation can take place, however, under a harmonic wave for which the line subtends an angle between 90 and 180 degrees and becomes positive or lagging.

From equations 23a

$$X_L = X_1\cot\theta_s \quad (24a)$$

$$\omega L = X_1\cot\theta_s \quad (24b)$$

But

$$\omega = 2\pi c/\lambda \quad (25a)$$

and

$$\lambda = 4s' = 4s(\pi/2)/\theta_s \quad (25b)$$

It follows that

$$\omega = c\theta_s/s \quad (25c)$$

which is the same as equation 10.

The substitution of equation 25c into equation 24b yields

$$X_1 s/cL = \theta_s/\cot\theta_s = \theta_s\tan\theta_s = L_s/L_0 \quad (26)$$

This equation is solved conveniently with the help of Figure 12.

From equation 20 we obtain for μ

$$\left.\begin{aligned} \mu_1 &= E_1/e_s \\ &= 1/\cos\theta_s \\ &= 1/\sin\theta_0 \end{aligned}\right\} \quad (27a)$$

and from equation 25b

$$s' = s(\pi/2)/\theta_s \quad (27c)$$

It may be observed that Figure 10 also may be considered as one-half of Figure 4, because the line of the latter could be cut in two without altering the oscillation of the system in any way. The foregoing equations differ from those of the closed line only because the origin of co-ordinates now corresponds to the middle of the line of Figure 4. The equations of the open and closed lines could be correlated along this idea, but no space will be devoted to it here.

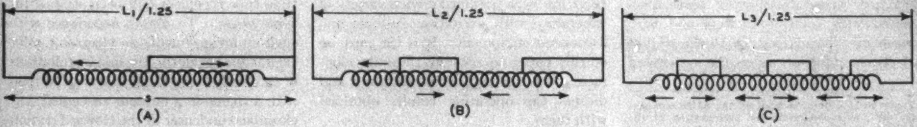


Figure 21. Connection to measure the inductance offered to various modes

A. The fundamental oscillation
B. The second harmonic oscillation
C. The third harmonic oscillation

Higher Harmonic Modes of the Open Line

Equations 20 and 21 apply to harmonic voltages and currents by substituting the appropriate value of λ. Therefore equations 24a and 26 which are derived from them also apply to the higher harmonics, Figure 13, and can be calculated with the help of the curves of Figure 12. When θ_s is determined for any harmonic, μ_n for it follows from equation 27b.

Appendix III. Other Combinations of Transmission Line and Massed Inductance

Although not all combinations of lines with massed inductances are capable of supporting standing waves, there is an enormous number that can. A few comments here on the general principles of these may be interesting and helpful.

1. For a stable standing-wave oscillation, the massed inductance must be adjacent to a voltage node (zero). Conversely, only those oscillations are possible as standing waves which bring the massed inductance to a voltage node.

2. As the current at both ends of a massed inductance has to be the same, such an inductance can be inserted into an interior nodal point only when the current and voltage of the inductance can be fitted to both sides of the circuit. For instance, a line open at both ends, Figure 14, with a massed inductance in the middle can oscillate in the manner shown. Figure 15 illustrates a possible case in which the massed inductance is located at approximately a quarter of the line and the oscillation is a full wave.

3. Figure 15 might be analyzed as a repetitive extension of Figure 11. In Figure 15, if s_1 is the short section of the line, s the total, and s_L the equivalent additional length of line introduced by L_0, then for a stable standing-wave oscillation like that shown, s_1 will not be $s/4$ but

$$s_1+(s_L/2)=(s+s_L)/4$$

Appendix IV. Effect of Massed Capacitance at Transformer Terminals

If one high voltage terminal is grounded, as assumed in the foregoing text, and the other not entirely isolated but connected to an idle network of appreciable capacitance, not large enough to be considered a short circuit to the transient phenomena, the situation is as if the winding is grounded through a capacitive reactance, and the foregoing equations are modified as follows.

Now i_x will not be zero at $x=s$ but some value i_s determined by

$$\left.\begin{aligned} i_s &= -e_s/X_c \\ &= -E_1(\sin\theta_s)/X_c \\ &= -I_{\max}(\sin\theta_s)(X_1/X_c) \end{aligned}\right\} \quad (42)$$

Equations 29a and 29b now become

$$i_x = i'_x+i''_x+i_s \quad (43a)$$

$$= I_{\max}\{\cos\,[(x/s)\theta_s]-\cos\theta_s-(\sin\theta_s)(X_1/X_c)\} \quad (43b)$$

Then

$$i_y = -(1/s)\int_0^s i_x dx \quad (44a)$$

$$= -I_{\max}\left[\frac{\sin\theta_s}{\theta_s}-\cos\theta_s-(\sin\theta_s)(X_1/X_c)\right] \quad (44b)$$

Equation 33a now becomes

$$E_1\sin\theta_s = (E_1/X_1)\left[\frac{\sin\theta_s}{\theta_s}-\cos\theta-(\sin\theta_s)(X_1/X_c)\right]\omega L_0 \quad (45a)$$

$$1 = (1/X_1)[1/\theta_s-\cot\theta_s-X_1/X_c]\omega L_0 \quad (45b)$$

The Development of Modern Excitation Systems for Synchronous Condensers and Generators

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DURING the past 15 years and particularly in the past ten years, there have been many major developments in excitation systems for synchronous condensers and generators. These changes and improvements were stimulated in 1928 by the introduction of the hydrogen-cooled synchronous condenser which necessitated a reconsideration of the excitation arrangement and brought about a change from a direct-connected exciter to separate excitation. This, together with the desire to obtain maximum speed of response and to eliminate maintenance caused by moving parts associated with rotating exciters and with mechanical regulators, led to the trial of electronic applications on the systems of the American Gas and Electric Company A development program along these lines was carried on over a period of approximately 15 years and included specifically 15 exciter applications on large generators and condensers. About six years ago a review of the performance and operating experience on the machines involved in this plan resulted in the design and trial of a system using a direct-connected exciter with amplidyne control for excitation of a large synchronous condenser. Following this, four similar installations were provided. Thus there has been obtained a wide variety of experience with excitation systems and associated equipment. It is the purpose of this paper to recount these developments, describe typical installations, and discuss the operating results obtained with them.

History of Development and Description of Installations

INTRODUCTION OF HYDROGEN-COOLED CONDENSERS

Until 1928 the conventional type of excitation system used with synchronous condensers consisted of a direct-connected or motor-driven exciter with mechanical regulator. Although satisfactory performance was, and still is, being obtained from rotating exciters, periodic shutdowns of the equipment were necessary for maintenance purposes. Also, the wear on the moving parts of the mechanical type voltage regulators necessitated frequent adjustments and replacements. In addition the quality of the performance of the regulating system was limited by the characteristics of the regulator and by the time lag of the exciter. At this time the first large hydrogen-cooled synchronous condenser, a 20,000-kva unit, was placed in service at the Turner station of the Appalachian Electric Power Company. The complication involved in attempting to build a gas tight machine with a direct-connected exciter discouraged the use of such an arrangement, and therefore a motor-driven exciter with a vibrating type regulator was provided.

In 1930 and 1931, four additional hydrogen-cooled synchronous condensers ranging in size from 15,000 to 30,000 kva were installed. This was discussed in a previous paper.[1] The excitation system for each of these machines likewise consisted of a motor generator set with mechanical type regulator.

ELECTRONIC PILOT EXCITERS

The possibility of taking advantage of high speed response characteristics of a tube type regulator had been studied for some years. To obtain experience with such an arrangement, an electronic pilot exciter and regulator, the first installations of this kind, were placed in service in 1930 with a 15,000-kva outdoor air-cooled synchronous condenser at the Howard station of the Ohio Power Company. The field of the motor-driven exciter of this unit is excited by means of a thyratron rectifier arrangement as described about ten years ago.[2] Pertinent data covering this machine and exciter are included in Table I, which lists all of the machines and excitation systems involved in this development program carried out on the systems of the American Gas and Electric Company.

Satisfactory performance obtained from this initial installation resulted in providing two similar thyratron pilot exciters and regulators for the direct-connected exciters of two large turbogenerators at the Glen Lyn station of the Appalachian Electric Power Company. These equip-

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The authors wish to acknowledge the fine spirit of co-operation and great amount of work carried out by the engineering staffs of both the General Electric Company and Westinghouse Electric Corporation, who between them designed and constructed all of the equipment discussed in this paper. The authors also express their indebtedness to the organizations of the Appalachian Electric Power Company, Indiana and Michigan Electric Company, Kentucky and West Virginia Power Company, Inc., Ohio Power Company, and the Scranton Electric Company, upon whose systems these machines were installed. Had it not been for the patient co-operation of all concerned in solving miscellaneous problems incident to the initial trials and later operation of the various excitation systems described in this paper, the development program covering this equipment could not have been brought to its unusually successful conclusion.

$$L_{s1}/L_s = 1-\theta_s/\tan\theta_s-\theta_s^2C/C_s \qquad (45c)$$

in which

C = the line capacitance connected to the transformer
C_s = the effective winding capacitance (analogous to L_s) for the harmonic under consideration

These two constants L_s and C_s are related to each other by the X_1 and the ω of the winding for the harmonic

$$X_n=\sqrt{L_{sn}/C_{sn}};\ \omega_n=\theta_{sn}/\sqrt{L_{sn}\times C_{sn}} \qquad (46)$$

If X_n, ω_n, and θ_{sn} are obtained by test, L_{sn} and C_{sn} can be calculated.

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3. LIGHTNING SURGES TRANSFERRED FROM ONE CIRCUIT TO ANOTHER THROUGH TRANSFORMERS, P. L. Bellaschi. AIEE TRANSACTIONS, volume 62, 1943, December section, pages 731–8.

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ments were placed in service in April 1931 and were installed primarily to prevent the occurrence of out-of-step conditions following severe faults. This is accomplished by employing a regulating system which maintains a steady bus voltage by controlling it continuously. This arrangement is provided with a grid-controlled 3-phase thyratron rectifier in place of a separate pilot exciter. The thyratron grids are energized by a 60-cycle voltage which lags the anode voltage by 90 degrees, and on this voltage is superimposed the variable direct voltage output of an electronic regulator.

Figure 1 shows the electronic regulator panel at Glen Lyn and Figure 2 is a schematic diagram of the circuit employed which is typical for both exciters as well as for the one at Howard. The diagram does not include filter circuits which are not essential to an understanding of the principles involved. The regulator employs a Wheatstone bridge having two legs of thyrite and two of standard resistors. The 3-phase voltage to be regulated is rectified and impressed on two terminals of this bridge. Because of the characteristics of thyrite, small changes in this impressed voltage produce relatively large changes in the voltage across the other two bridge terminals. This latter voltage is applied to the grid of a pliotron, producing an amplified voltage across a resistor in the pliotron plate circuit. This amplified voltage is used as the variable d-c grid control voltage for the thyratron rectifier. The thyratron tube heaters are supplied from the same potential transformers which supply the regulator. A means is provided for transferring the main exciter from electronic pilot excitation to self-excitation when necessary to change tubes or to work on the regulator pilot exciter system.

ELECTRONIC MAIN EXCITERS

The desire to obtain the maximum benefit from an electronic system following the satisfactory performance of the tube type pilot exciters previously mentioned resulted in the installation in August 1933 of the first all tube regulator and thyratron excitation system at Scranton, Pa. This equipment was placed in service with a 15,000-kva hydrogen-cooled synchronous condenser and has been described elsewhere.[3] In the original installation the six power tubes were air-cooled thyratrons, but they were replaced several years later with water-cooled ignitrons.

A similar combined electronic exciter and regulator of improved type were installed at Fostoria, Ohio, on a 36,000-kva synchronous condenser in 1936.[1] The outstanding feature of this machine was its design for operation at a gas pressure of 15 pounds per square inch, utilizing for the first time the advantage of more efficient cooling, obtainable by operating at higher gas pressure.

Satisfactory performance of the electronic exciter at Fostoria as well as the one at Scranton led to the installation of a third exciter for a similar 36,000-kva synchronous condenser at the Torrey Station of The Ohio Power Company. This equipment was provided in 1939 for maintaining the voltage of a 22-kv bus which serves a highly fluctuating load consisting primarily of large electric furnaces. Figure 3 shows a front view of the Torrey electronic excitation equipment which is mounted on the three left panels of the switchboard. The tube compartments were designed as an integral part of this switchboard, two tubes being mounted behind each panel. Access for changing tubes is from the rear and the anode circuit breaker must be tripped before the compartment door can be opened. Furthermore, the main field circuit which is normally ungrounded must be tested for grounds, and if it is found clear the cathode bus must then be

Table I. Typical Synchronous Condensers and Generators Associated With Development of Excitation Systems

Listing as of February 1, 1946

			Synchronous Machine				Main Exciter			
								Rating		
Location	Operating Company	Starting Date	Type	Kilovolt-Amperes	Revolutions Per Minute	Cooling Medium	Type	Kilowatts	Volts	Field Supply or Control
Howard, Ohio	The Ohio Power Company	March 1930	Condenser	15,000	900	Air	Motor driven	80	250	Electronic pilot exciter
Glen Lyn, Va.	Appalachian Electric Power Company	*April 1931	Turbogenerator	25,000	1,800	Air	Direct connected	100	250	Electronic pilot exciter
Glen Lyn, Va.	Appalachian Electric Power Company	*April 1931	Turbogenerator	31,250	1,800	Air	Direct connected	150	250	Electronic pilot exciter
Scranton, Pa.	The Scranton Electric Company	August 1933	Condenser	15,000	900	Hydrogen	Electronic	78	150	
Fostoria, Ohio	The Ohio Power Company	November 1936	Condenser	36,000	1,200	Hydrogen	Electronic	175	200	
Fostoria, Ohio	The Ohio Power Company	*February 1938	Condenser	5,000	600	Air	Direct connected	54	125	Electronic pilot exciter
Torrey, Ohio	The Ohio Power Company	July 1930	Condenser	36,000	1,200	Hydrogen	Electronic	175	200	
Claytor, Va.	Appalachian Electric Power Company	October 1939	Hydrogenerator	20,833	138.5	Air	Direct connected	150	250	Electronic pilot exciter
Claytor, Va.	Appalachian Electric Power Company	October 1930	Hydrogenerator	20,833	138.5	Air	Direct connected	150	250	Self-excited amplidyne control
New Carlisle, Ind.	Indiana and Michigan Electric Company	March 1940	Condenser	30,000	3,600	Hydrogen	Electronic	125	250	
Muncie, Ind.	Indiana and Michigan Electric Company	May 1941	Condenser	18,750	900	Hydrogen	Direct connected	120	250	Self-excited amplidyne control
Danville, Va.	Appalachian Electric Power Company	August 1942	Condenser	18,750	900	Hydrogen	Direct connected	120	250	Self-excited amplidyne control
Lima, Ohio	The Ohio Power Company	October 1943	Condenser	18,750	900	Hydrogen	Direct connected	120	250	Self-excited amplidyne control
Beaver Creek, Ky.	Kentucky and West Virginia Power Company, Inc.	January 1945	Condenser	18,750	900	Hydrogen	Direct connected	120	250	Self-excited amplidyne control
Anderson, Ind.	Indiana and Michigan Electric Company	January 1946	Condenser	18,750	900	Hydrogen	Direct connected	120	250	Self-excited amplidyne control

* Electronic pilot exciters placed in service on this date on machines installed previously.

Figure 1. Back view of electronic pilot exciter and regulator

Glen Lyn 31,250-kva generator of the Appalachian Electric Power Company

grounded before the door can be opened. With these precautions tubes may be changed without shutting down the machine. A means is provided for cutting out the electronic regulator and leaving the exciter itself on manual control.

Figure 4 is a schematic diagram showing the circuit employed for this unit. As in the case of the pilot exciters, all filter circuits not essential to an understanding of the fundamentals of the control are omitted. The rectifier is 6-phase and the point at which the igniters fire is controlled by six thyratrons, the grids of which are energized from a grid transformer which supplies a voltage lagging the anode voltage by 120 degrees. On this voltage is superimposed a variable d-c output voltage from the electronic regulator in a manner similar to that employed for the electronic pilot exciters. Each power tube has an individual anode circuit breaker provided with reverse current protection.

The regulator employs a glow tube which provides a constant voltage against which variations in regulated voltage are compared. The 3-phase voltage to be regulated is rectified and impressed on the glow tube through a resistor. The voltage across the glow tube provides a constant potential for the cathode of a pliotron tube whose control grid voltage is obtained from a potentiometer connected across the rectified regulated voltage. Small changes in regulated voltage therefore produce changes in control grid voltage, which are amplified by the pliotron, producing large changes in the voltage across a resistor in its plate circuit. As in the case of the bridge-type regulator, this amplified voltage is applied to the thyratron firing tube grids to control their output and the output of the ignitron rectifier. Although omitted from the diagram for simplicity, all tubes in the regulator are provided in duplicate and connected in parallel so as to provide protection against tube failures.

A fourth installation of an electronic main exciter and electronic voltage regulator was completed in 1940 on a 30,000-kva synchronous condenser at New Carlisle, Ind. As reported elsewhere,[1] this condenser was designed to take maximum advantage of the low density of hydrogen gas by using for the first time a speed of 3,600 rpm. This rating of the machine is based on operation at a gas pressure of 15 pounds per square inch. Figure 5 shows a front view of the exciter installation. The three panels at the left of the switchboard are used for the control of the condenser and its switching. As in the case of Torrey, the power tubes may be changed without shutting down the machine. The regulator which supplies variable d-c bias for the thyratron tube control works on a somewhat different principle from that of Torrey. The control grid voltage for the pliotron tube is made up of two components. The first is a rectified voltage which varies directly with the regulated voltage and which therefore acts similarly to the Torrey arrangement. The second is a voltage drop across a hot cathode tube which varies with the filament voltage. The filament voltage of this tube is obtained from the regulated voltage. A change in regulated voltage therefore produces an instantaneous change in the pliotron control grid voltage, followed by a somewhat slower further change as the tube filament temperature changes. This arrangement is designed to keep hunting at a minimum

Miscellaneous Direct-Connected Exciter Systems

In October 1939, four 20,833-kva hydrogenerators were placed in service at the Claytor station of the Appalachian Electric Power Company. All of these machines were provided with direct-connected exciters, but in order to obtain comparative experience on various types of pilot exciters and regulators, two of the machines were supplied with direct-connected pilot exciters and rheostatic type voltage regulators, and one was equipped with an electronic pilot exciter similar to those previously installed at Glen Lyn and Howard and with a regulator similar to that at Torrey. An amplidyne excitation system with a static type voltage regulator was provided for controlling excitation of the fourth machine. This was the first time that an amplidyne excitation system had been provided for a syn-

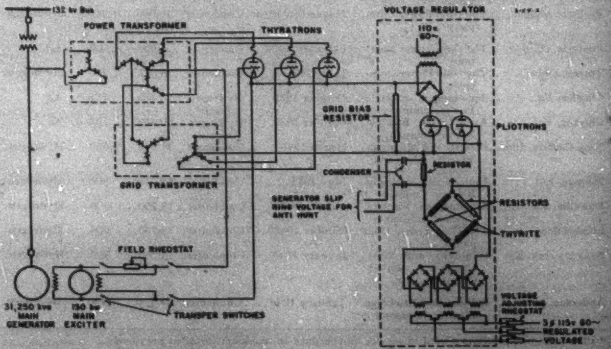

Figure 2. Schematic diagram of electronic pilot exciter and regulator

Glen Lyn 31,250-kva generator

Figure 3. Front view of electronic main exciter and regulator

Torrey 36,000-kva synchronous condenser of The Ohio Power Company

chronous generator. An explanation of this type of arrangement is included later.

During the period from 1930 to 1940, other waterwheel generators and turboalternators were installed which were provided with the customary arrangement of direct-connected or gear-driven exciter and pilot exciter. Also air-cooled synchronous condensers in capacities up to 7,500-kva were placed in service at various locations and these were provided with direct-connected exciters and improved types of mechanically operated voltage regulators. During this time there was considerable development on regulator applications for the purpose of obtaining greater reliability and better performance.

Review of Development Up to 1940

In 1940 the performance record of, and operating experience obtained with, all of the synchronous condensers and generators and their excitation and regulation systems previously described were reviewed in order to determine whether definite conclusions could be reached as to an ideal arrangement. The excellent performance anticipated from the electronic exciter facilities was demonstrated without question. However, the initial cost of such equipment in addition to the heavy expense of making tube replacements offered objections to continuing this arrangement. The superior performance and reliability of the electronic regulator definitely had been proved. Difficulties resulting from operating synchronous condensers at speeds of 1,200 and 3,600 rpm indicated the desirability of returning to the previously conventional speed of 900 rpm. The use of hydrogen at gas pressures up to 15 pounds per square inch was established definitely as being a highly satisfactory and desirable practice, and therefore consideration was given to the design of a machine suitable for still higher gas pressure operation. Also, the possibility of providing a direct-connected exciter for a hydrogen-cooled unit had been given considerable study by the manufacturers, and there was reason to believe that a satisfactory design could be made. The superiority of a direct-connected arrangement, as far as electrical performance is concerned, could be obtained in this way. Furthermore, such a design would offer the many advantages of hydrogen atmosphere operation for the exciter as well as for the condenser. The amplidyne excitation arrangement at Claytor was providing excellent performance and therefore was given due consideration in the final decision for obtaining the most suitable excitation arrangement for a synchronous condenser.

Direct-Connected Exciters With Amplidyne Control

As a result of this careful survey of the whole problem, an entirely new type synchronous condenser running at 900 rpm was designed by one of the manufacturers and placed in service in 1941 at the Muncie station of the Indiana and Michigan Electric Company. This is a hydrogen-cooled machine and its rating at one-half pound-per-square-inch hydrogen pressure is 18,750 kva. It was designed for operation at gas pressures up to 25 pounds per square inch and a maximum capability of approximately 25,000 kva is obtained on this basis. This capability is realized partly because of the advantages of operating at the increased hydrogen pressure and partly because of the margins which exist in the machine at the one-half pound rating. The outstanding development in connection with this unit is that for the first time a direct-connected exciter is provided in a hydrogen ventilated machine. The normal air-cooled rating of this exciter is approximately 100 kw but by utilizing the efficient cooling effect of hydrogen at a gas pressure of 25 pounds per square inch, the capacity is increased to approximately 150 kw. The amplidyne regulation system was provided for controlling the excitation to the condenser field and an electronic regulator was installed for controlling the amplidyne generator. Thus an ideal arrangement for an excitation system for a synchronous condenser is obtained when consideration is given to the various factors involved such as greatest reliability, best performance, highest economy, and least maintenance. Along with this improvement there evolved a superior design for the condenser itself.

A schematic diagram showing the electrical connections for this installation is shown in Figure 6. Excitation is controlled by the amplidyne generator which is connected as a source of voltage in series with the field of the main exciter, and which is provided with two control fields. Its buck field is supplied with constant excitation by a small rectifier while its boost field is supplied by the same rectifier through a pliotron tube, the grid of which is controlled by the output of an

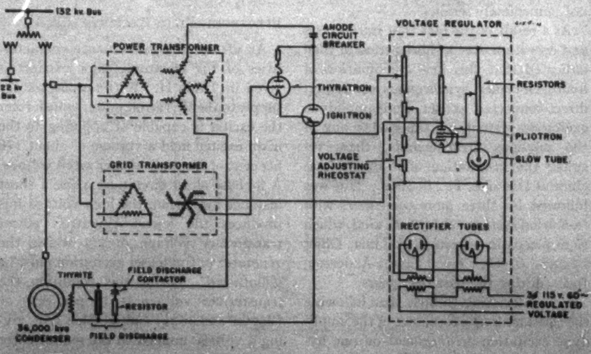

Figure 4. Schematic diagram of electronic main exciter and regulator

Torrey 36,000-kva synchronous condenser

Figure 5. Front view of electronic main exciter and regulator

New Carlisle 30,000-kva synchronous condenser of The Indiana and Michigan Electric Company

electronic voltage standard type regulator similar to that shown in Figure 3 for Torrey. Thus the amplidyne serves as the final step of amplification of the variations in regulated voltage and provides the necessary power to change the field current of the main exciter quickly. The amplidyne driving motor power source is taken from the station auxiliary power supply. It may be noted from Figure 6 that a fixed resistance is connected in series with the amplidyne. This resistance limits the voltage which the main exciter will attain as a self-excited exciter to a value only slightly in excess of normal voltage, and therefore makes accidental loss of amplidyne motor driving power not immediately serious.

As a result of the fact that initial tests and operation over a short period of time indicated that this new combination of hydrogen-cooled synchronous condenser, direct-connected exciter, and amplidyne excitation system was superior to any of the previous arrangements, there resulted the installation of a similar machine at Danville, Va., in 1942. This was followed by three more condensers with excitation systems of the same kind, which were placed in operation at Lima, Ohio; Beaver Creek, Ky.; and Anderson, Ind.; during the past four years.

It is interesting to note that following the satisfactory performance of the amplidyne excitation arrangement on one hydrogenerator and five condensers over the past six years, the same system is being utilized for a 32,500-kva turbogenerator which will go into service this year at Atlantic City, N. J. A similar arrangement is being provided for another 32,500-kva turbogenerator now being built for installation at Scranton, Pa.

Operating Experience and Performance Data

ELECTRONIC PILOT EXCITERS

An idea of the performance characteristics of the electronic pilot exciters is given in Table II. With normal voltage supply to the thyratron power transformer the exciter is capable of applying to the main exciter field a voltage of about 170 per cent of the main exciter rated voltage. A 3-phase short circuit on a line a short distance away from the high voltage bus produces the lowest thyratron power transformer voltage during which the generator or condenser excitation must be maintained. Under this condition the transformer voltage is 14 per cent at Claytor and 35 per cent at Glen Lyn, giving a voltage applied to the main exciter field of 25 per cent at Claytor and 60 per cent at Glen Lyn. This is equivalent to about full load excitation in the case of Claytor. There have been no cases of loss of synchronism resulting from excitation failure at either plant.

It must be borne in mind that 3-phase faults are comparatively rare and, in addition, that such a fault on any line at either station will be cleared in a maximum of nine cycles. Furthermore, at Claytor it is possible to operate two machines simultaneously connected to the low voltage side of the transformer bank and this is a normal operating condition whenever the stream flow permits. In this case the maximum pilot exciter output voltage would be about 50 per cent of the main exciter voltage rating. Likewise at Howard where the maximum instantaneous pilot exciter output voltage for a 3-phase high voltage fault is 30 per cent of the main exciter rating, there has been no indication of loss of synchronism. In this case where the machine is a synchronous condenser, momentary loss of synchronism might go undetected because there is no out-of-step relay for protection.

Table II shows the types of thyratron tubes employed and the life obtained from them. At Howard it can be noted that two types of tubes have been used. The original tubes were chosen conservatively to have about five times as much current carrying capacity as was required.

Table II. Application Data and Performance Record of Electronic Pilot Exciters

Listing as of January 1, 1946

Location	Per Unit Reactance: Generator or Condenser X''_d	Per Unit Reactance: Main Transformer X	Pilot Exciter Max Output Volts	Thyratron Tubes: Type	Thyratron Tubes: Rated Plate Amp	Total Tube Failures	Average Operating Hours Per Tube Failed	Tota Operating Days of Pilot Exciter
Howard (original)	0.226	0.048	425	FG-29	12.5	13	9,998	1,745
Howard (modernized)	0.226	0.048	425	FG-95	2.5	19	13,632	3,650
Glen Lyn units 2 and 4 (original)	0.16	0.086	425	FG-29	12.5	18	23,016	7,625
Glen Lyn units 2 and 4 (modernized)	0.16	0.086	425	FG-172	6.4	5	12,613	
Fostoria	0.20	0.029	160	FG-172	6.4	9	10,033	2,942
Claytor	0.271	0.044	440	FG-172	6.4	24	3,376	1,208
Average							11,735	
Total						88		17,170

The tubes now used have a current rating about equal to the exciter field requirements. In spite of this difference the original tubes did not give as good service as the new ones. However, these original tubes were subjected to more severe service because of the type of control circuit used which caused unbalanced tube currents with relatively small changes in grid characteristics. A similar change in tubes but not in control was made at Glen Lyn, and in this case the performance of the original tubes appears to be almost twice as good as that of the new ones. However, this change has been made quite recently, and it will be noted that only five of the new tubes have failed so far.

At both Glen Lyn and Howard the tube life record for the original installation includes the operating hours of the tubes which were in service but had not failed when the exciter was modernized. The tubes now in service at Glen Lyn have an average operating record of over 17,000 hours, and it therefore appears that the average operating hours of the tubes which fail will increase considerably as time goes on. The record for the thyratrons of the Claytor exciter is not as satisfactory as that for the other machines. The exciter field ampere requirements are well within the tube ratings and so far no explanation has been found for this relatively shorter life.

Aside from expense for replacement of thyratron tubes, the maintenance cost of these exciters has been negligible. Because tubes may be installed without shutting down the machine, tube replacement does not represent an operating hardship. From the data in Table II it will be noted that the average operating time per tube for all installations has been 11,735 hours. There have been about 2.2 tubes required per exciter for each year of actual operation and it is expected that further operation will reduce this figure. Total combined operating time for all electronic pilot exciters as of January 1, 1946 was 17,170 days or approximately 47 years.

ELECTRONIC MAIN EXCITERS

The performance characteristics of electronic main exciters are given in Table III. With normal voltage supply to the excitation power transformer the exciter has a ceiling voltage of 275 per cent of maximum continuous slip-ring voltage rating in the case of Torrey, Fostoria, and New Carlisle; and of 244 per cent in the case of Scranton. When a 3-phase short circuit occurs on the high voltage bus, the output voltage varies from 127 per cent of slip-ring voltage in the case of New Carlisle to 41 per cent in the case of Scranton. As in the case of the electronic pilot exciters, there has been no indication of loss of synchronism on any of the machines equipped with this type of exciter. Figure 7 shows regulated voltage of the 22-kv bus and condenser loading for a portion of a typical day's operation at Torrey. The 36,000-kva condenser is connected through a 3-winding 48,000-kva transformer bank to the 132- and 22-kv station busses as shown in Figure 4. The condenser load may be seen to vary between 40,000 kilovar out and 11,000 kilovar in, the latter being the zero field current load, with variations in regulated voltage of less than one per cent. These charts are typical of those obtained from the other machines. The Torrey unit was chosen for illustration because it is subjected to frequent large changes of load caused by the fluctuating nature of the arc furnace load served from the 22-kv bus.

Table III shows the types of thyratron and ignitron tubes employed and the tube

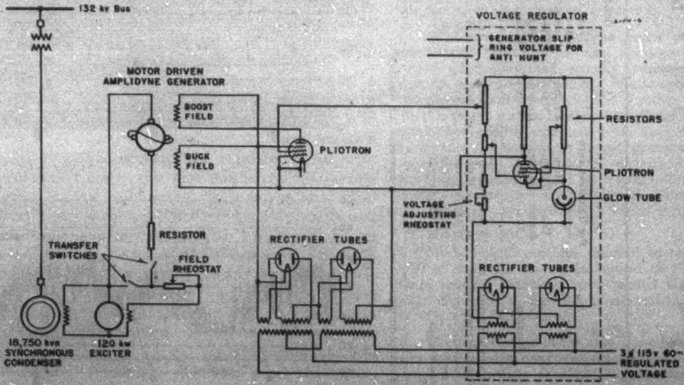

Figure 6. Schematic diagram of direct connected exciter with amplidyne control

Muncie 18,750-kva synchronous condenser of The Indiana and Michigan Electric Company

Table III. Application Data and Performance Record of Electronic Main Exciters

Listing as of January 1, 1946

Location	Per Unit Reactance		Exciter Max Output Volts	Condenser		Thyratron Firing Tubes				Power Tubes					Total Operating Days of Exciter
	Condenser X"d	Main Transformer X		Max Continuous Slip-Ring Volts	Max Continuous Field Amps	Type	Rated Plate Amp	Tubes Failed	Avg Operating Hrs Per Tube Failed	Type	Rated Plate Amp	Tubes Failed	*Avg Operating Hrs Per Tube Failed	Avg Operating Hrs of Tubes in Service	
Scranton (original)	0.25	0.05	280	115	475	FG-95	2.5			FG-53	100	10	22,029	17,683	3,494
Scranton (modernized)	0.25	0.05	280	115	475	FG-172	6.4	16	8,651	FG-259	100	8	18,045		
Fostoria	0.32	0.076	435	160	740	FG-172	6.4	23	13,277	FG-238	150	16	12,940	19,651	3,021
Torrey	0.32	0.088	435	160	740	FG-172	6.4	14	17,678	FG-238	150	8	25,736	15,563	2,310
New Carlisle	0.100	0.093	580	217	500	KU-676	6.4	20	9,314	WL-653	150	3	5,900	36,380	1,917
								73 (Total)	12,021 (Avg)			45 (Total)	16,340 (Avg)		10,742 (Total)

* Failures during first year of exciter operation are omitted.

life obtained from them. In the case of the power tubes, failures during the first year of operation have been omitted because in each case this proved to be a trial period during which various modifications were made to improve performance. In the case of the firing thyratrons, the entire life of the exciter is included because the initial trial period made little difference in the averages. In order to indicate the trend of averages with further operation, the average operating hours of the power tubes now in service have been included. These figures indicate that much longer tube life is to be expected both at Fostoria and New Carlisle where the tube designs have been improved and the tube failure hours have been influenced by the record of those of earlier design as well as by a few defective tubes which failed with comparatively short service records. The record for the original installation at Scranton includes the tubes which were in service but had not failed when the exciter was modernized.

As in the case with the pilot exciters, the maintenance cost of these exciters has been negligible with the exception of the expense for tube replacements. From the data in Table III it will be noted that the average operating hours per power tube is 16,340 and per firing tube is 12,021. Since six tubes of each type are employed per exciter there have been 3 power tubes and 4.4 firing tubes required for each year of actual operation, and it appears that further operation should produce a substantial improvement in the record for the power tubes. The total combined operating time for the four electronic main exciters as of January 1, 1946 was 10,742 days or approximately 29 years.

Amplidyne Excitation Systems

The amplidyne excitation system performance has been similar to that obtained from the electronic pilot and main exciters. From the standpoint of power system operation, no essential difference has been observed on the condenser installations between the performance of excitation systems with ultrahigh speed response and those having more conventional speeds. For small changes in regulated voltage the speed of response of the amplidyne system is comparable to that of the electronic pilot exciter. This is true because the controlling factor in both cases is the amount of antihunt which must be introduced to produce stable operation. The ceiling voltage of the amplidyne generator is not affected by large dips in system voltage because its driving motor is designed with considerably more power than normally required and, in addition, its inertia contributes to the total energy required for quick changes in excitation. Figure 8 is an oscillogram of a special test made at Muncie in June 1941, and shows the rate of response of the amplidyne generator to a sudden change of $2^{1}/_{2}$ per cent in regulated voltage starting with a slip-ring voltage of 45. The measured maximum rate of response of slip-ring voltage as obtained

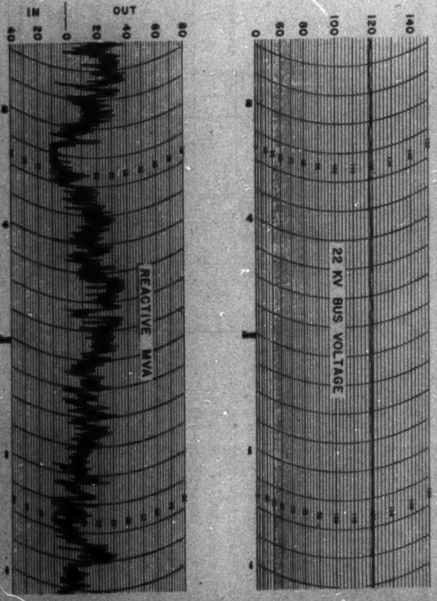

Figure 7. Typical charts of regulated voltage and condenser load

Torrey 36,000-kva synchronous condenser of The Ohio Power Company

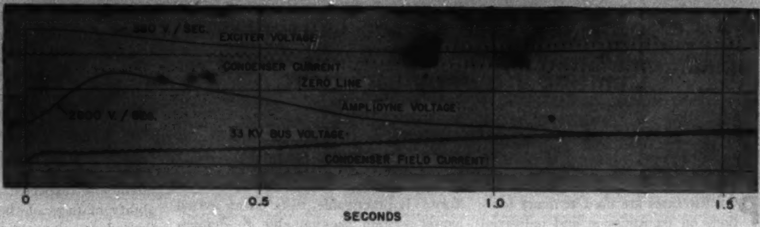

Figure 8. Oscillogram showing response of amplidyne control system

Muncie 18,750-kva synchronous condenser of The Indiana and Michigan Electric Company

from this oscillogram is 380 volts per second. The maximum rate of response of the amplidyne generator voltage was 2,800 volts per second.

Because the amplidyne generator can produce either a buck or a boost voltage, it is suited ideally to produce a reversed slip-ring voltage by which additional lagging capacity can be obtained from the condenser. This feature is made use of in all cases and gives an increase in capacity of about 18 per cent above that obtained with zero slip-ring voltage, without approaching the point at which the condenser would slip a pole. This feature cannot be obtained with an electronic exciter and can be obtained with an electronic pilot exciter only by the addition of a differential field and some source of excitation for it.

Maintenance of an amplidyne excitation system is essentially the same as for other arrangements using rotating exciters. The amplidyne generator driven by an ordinary induction motor requires very little attention. The direct-connected exciter operating in hydrogen has given very satisfactory results. The absence of external dirt and oxygen in the cooling gas for the exciter compartment has simplified the maintenance problem. Periodic cleaning and inspection of the exciter compartments is made three times a year, with the total annual outage for this reason amounting to approximately one day. As of January 1, 1946, these five machines have operated for a combined total time of 3,994 days or approximately 11 years.

Electronic Regulators

The performance of both types of electronic regulators described has been excellent. Both are capable of holding regulated voltage within about three-quarters of one per cent while varying the condenser loading from maximum kilovars in to rated kilovars out. The bridge type is somewhat more sensitive to large changes in ambient temperature, but as all are installed in heated control rooms, this is of little practical importance. The tubes used in these regulators have given long life and their replacement cost is low. As stated, charts of regulated voltage and condenser loading are similar for all the excitation systems described, and this is the case because all are controlled by sensitive electronic regulators which have no dead band and respond to the slightest change in regulated voltage. Aside from tube changes, only a few minor replacements of parts have been required. In two cases copper oxide rectifiers have been replaced and in one case a voltage adjusting rheostat of the conventional slide wire type was changed.

The electronic regulator lends itself readily to the addition of special features when desired. Some of those which have been employed are as follows:

1. A definite slip-ring voltage set automatically for transferring the condenser from starting to running.
2. A definite maximum slip-ring voltage set automatically when the condenser approaches a dangerous temperature.
3. Adjustable amounts of droop obtained in regulated voltage.
4. Adjustable maximum reverse slip-ring voltage used in the case of amplidynes.

Its accuracy and the fact that it has no dead band make the electronic regulator readily adaptable to parallel operation in that compensator or droop settings can be made with minimum attention to errors in the regulator itself.

However, the primary purpose of a voltage regulator when used with a synchronous condenser or generator is to control voltage, and the more sensitive and accurate the regulator is the more nearly this aim can be realized. From a maintenance viewpoint an ideal regulator would have no moving parts and require no periodic adjustments or renewal of parts. Although the electronic regulator does require periodic replacement of tubes its absence of moving parts brings maintenance attention to a much lower level than is the case with any of the various types of mechanical regulators. The total combined operating time as of January 1, 1946 for the 14 electronic regulators installed on the exciters previously discussed amounts to approximately 87 years.

Conclusions

Operating experience with 5 electronic pilot exciters, 4 electronic main exciters, 14 electronic regulators, and 6 amplidyne excitation systems on 15 large synchronous generator and condenser applications, dating as far back as 1930, permits the drawing of rather definite conclusions.

1. More than 47 machine years of operation of electronic pilot exciters have demonstrated the reliability and satisfactory performance of this type of excitation when used with either generators or condensers on a large power system.
2. Operating experience with four electronic main exciters over a combined period of approximately 29 years has demonstrated the excellent performance and reliability of this excitation arrangement.
3. Initial cost of electronic exciter equipment and the heavy expense of making tube replacements offer objections to use of this otherwise most desirable equipment. It is hoped that further developments in the design of more satisfactory tubes for this purpose will occur which will reduce original and maintenance costs thus making the electronic system more attractive.
4. Approximately 11 machine years of satisfactory operation of five hydrogen-cooled direct-connected exciters indicates that the many benefits and fine results now being obtained by the operation of numerous generators and condensers in a hydrogen atmosphere can be duplicated in the case of exciters. The feasibility of gas pressure operation up to 25 pounds per square inch has been demonstrated. This results in increased capacity for both the exciter and the synchronous condenser to which it is connected.
5. More than 87 machine years of operation of electronic regulators applied on 14 large rotating machines have proved the reliability and the desirable characteristics of this method of control.
6. The performance of the amplidyne control system with direct-connected exciters has been essentially equal to that obtained

Performance Criteria for Current-Limiting Power Fuses—I*

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POWER fuses are used at many locations on transmission and industrial power systems to insure the prompt interruption of serious short-circuit currents. An engineer making such an application primarily desires assurance that the fuses will be capable of safely interrupting all magnitudes of short-circuit current up to the maximum the system can produce.

This available short-circuit interrupting capacity may be expressed either in terms of 3-phase symmetrical kilovolt-amperes, or as maximum rms current. In computing the latter, the initial symmetrical alternating component of current is multiplied by the factor 1.6 to provide for the maximum practicable d-c component which may be involved in the short circuit.

Either or both of these ways of expressing the interrupting requirements may be compared with manufacturers' interrupting ratings for power fuses. The maximum current method of rating has been in use longer, but in recent years increasing use has been made of the 3-phase kilovolt-ampere rating, to facilitate direct comparison with the ratings given circuit breakers. In making reference to such ratings, the user is relying on the manufacturer to have made adequate and sufficiently severe tests on the fuse units to insure that they will interrupt safely fault current such as may occur within their interrupting ratings.

The proving of the interrupting ratings of power fuses usually is done in a short-circuit testing laboratory, where all factors affecting the test severity can be controlled and measured. Figure 1 shows power fuses arranged for test in the short-circuit cell of a high power testing laboratory. If practicable, tests on an actual power system are valuable to demonstrate the ability of the fuses to interrupt fault current at the location where their use is proposed. Such tests are likely to be less severe than those made in a short-circuit testing laboratory because of the presence of line resistance, and because the operating voltage usually will be less than the voltage rating of the fuse unit for the same voltage class. At the laboratory, it is also possible to vary all factors contributing to test severity, and to measure their separate effects on fuse performance. One important device available at the testing laboratory is the synchronous closing switch, with which it is possible to explore the regions of fault starting angle which analysis indicates might be severe. Such facilities permit the rapid accumulation of useful data, which could be obtained only with considerably greater difficulty by making tests at random closing angles, as would be necessary on a power system where synchronous closing means are not available. With laboratory facilities, the fuse designer can carry interruption tests to the point of destruction of his developmental fuse samples to disclose critical design factors. It is necessary to have such data to design power fuses having a factor of safety above the required interrupting ability.

Modern power fuses are of two principal types; those that permit approximately the full available short-circuit current to flow during the first half cycle, and those which have a substantial current-limiting effect.

The latter is the more recently developed type[2] and is a more efficient interrupter because the considerably smaller energy dissipated within the fuse unit can be absorbed there without external disturbance. In fuses which do not have this current limiting effect, the arc energy released in the fuse is usually so high that it must be expelled to the outer atmosphere or into a chamber where the ionized gases can be condensed. As might be expected, the interrupting ability of fuses of the expulsion type is limited as compared to the interrupting ability of fuses of the current-limiting type. Figure 2 shows oscillograms taken during interruptions by both these types. Symbols have been added to the oscillograms appearing in this portion of the paper to assist in their interpretation. These symbols are defined in Appendix I, and are defined and used similarly in the analytical study presented in part II of this paper.[1]

Of the two types, the current-limiting fuse lends itself better to the study of severity. This results primarily from the prompt action in interrupting the available current, and from the fact that an oscillogram taken during the test will show very definitely whether the current was interrupted easily or with difficulty.

Increasing severity can be detected both by oscillographic records and by the visible evidence of distress shown by the fuse test sample. Figure 3 shows four test samples of current-limiting power fuse units after tests at the same voltage, but with progressively increasing available short-circuit current. In these, the excess filler has been removed, showing the "fulgurites" or fused tubes of granular quartz where the elements were before they were melted. These formations seemingly have become heavier as severity increased, and in the case of sample 4,

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*Part II of this paper[1] was prepared by E. W. Boehne.

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from the electronic main or electronic pilot applications.

7. The evolution of an ideal combination consisting of a standard speed synchronous condenser with direct-connected exciter operating in a hydrogen atmosphere at 25 pounds gas pressure and with amplidyne control and electronic regulator, illustrates the interdependence of a machine and its associated equipment and shows what can be accomplished by a co-ordinated development program over a period of years.

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enough damage was done to the interior of the sample by the arcing to cause the tube to break. Differences in test severity such as those between samples 1, 2, and 3 cannot be judged accurately from the remains of test samples, but oscillograms of the tests give dependable comparisons. For this reason, oscillograms of power fuse tests are included in part I of this paper to illustrate differences in severity resulting from changes in the variable circuit factors.

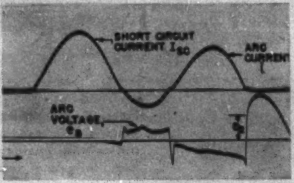
The severity of an interruption test on a power fuse depends on the following factors or variables of the test circuit:

1. Generated voltage.
2. Available current.
3. Fault starting angle.
4. Voltage angle at melting.
5. Frequency.
6. Power factor.
7. Number of phases.

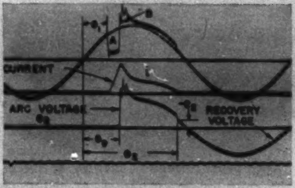
It is the purpose of this portion of the paper (part I), to appraise the separate effect on severity of each of these variables, with the object of arriving at recommendations for the testing of power fuses.

Generated Voltage

The circuit voltage at which interruption tests are made directly affects test severity, other factors remaining constant, as will be noted from the oscillograms shown in Figure 4. These record three tests on the same developmental design of test sample with progressive increases in circuit voltage. At 110 per cent of rated voltage (Figure 4A), the fuse had no difficulty in developing enough arc voltage drop to more than exceed the generated voltage and bring the current down to zero within one-quarter of a cycle. The arc current, i, was substantially constant during the first portion of the arcing period and then was forced down to zero without apparent difficulty. At 115 per cent voltage (Figure 4B) the fuse apparently had greater difficulty in interrupting the circuit because the arc current reached a higher value than the melting current, and lasted somewhat longer than at 110 per cent voltage. In both these tests the fuse interrupted with no external disturbance.



A. Expulsion fuse



B. Current-limiting fuse

Figure 2. Typical oscillograms of short-circuit interruptions by the two principal types of power fuses

At 120 per cent voltage (Figure 4C) the arc voltage drop developed by the fuse was insufficient to exceed this higher test circuit voltage, and after an initial dip the arc current i was permitted to rise to substantially the available short-circuit values. The resulting high release of energy within the fuse test sample caused the tube to break, arc current of short-circuit magnitude persisted, and the circuit had to be opened by the backup circuit breaker. In these tests, different degrees of severity were imposed upon a fuse unit of given design capable of developing only a given arc voltage drop. In part II of this paper,[1] the arc current is determined analytically, assuming certain ratios of fuse arc voltage to the circuit voltage, the results showing the same trend as the empirical results of Figure 4. These observations emphasize the importance of making interruption tests of power fuses at the rated voltage of the fuse units in accordance with established standards.

Available Current

Short-Circuit Current

As a result of the fact that the heat capacity of the current responsive element of a power fuse is relatively small, melting by a high short-circuit current can take place early in the first half cycle. The

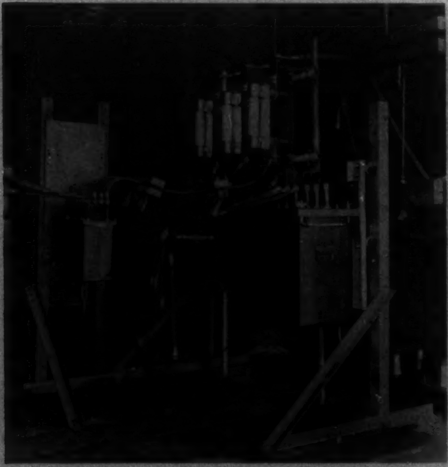
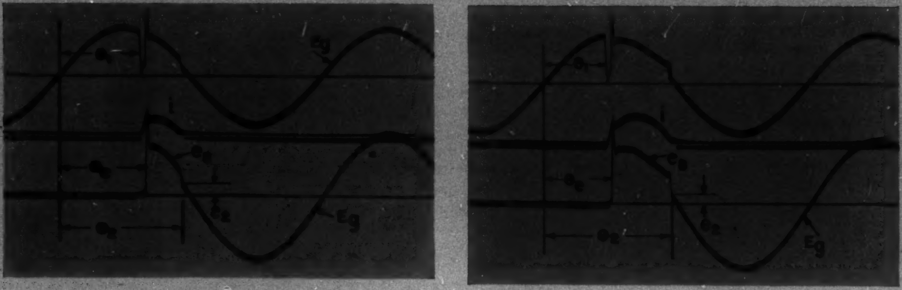


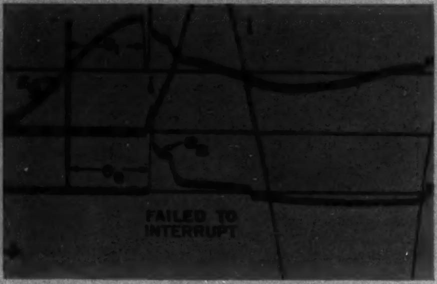
Figure 1 (left). Power fuses connected in circuit at short-circuit testing laboratory

Figure 3 (above). Current-limiting power fuse units after interruption tests of progressively greater severity



(A)

(B)



(C)

Figure 4. Illustrating progressive increase in test severity resulting from increasing test voltage

Developmental test samples

A. 110 per cent of rated voltage
B. 115 per cent of rated voltage
C. 120 per cent of rated voltage

melting time and the instantaneous value of current when melting takes place can be calculated with a high degree of accuracy.[1] The factors affecting the melting point are

1. The material and cross section of the current responsive element.
2. The frequency.
3. The available short-circuit current.
4. The degree of asymmetry of the available current.

The magnitude of an available symmetrical short-circuit current implies a certain initial rate of rise of current. The initial slope of the current-time curve is zero when considering a fully displaced asymmetrical current. For a symmetrical current, the peak current admitted by a fuse unit of low ampere rating varies directly as the cube root of the available short-circuit current, and for asymmetrical current, directly as the fifth root. For larger ampere ratings, a more involved calculation is necessary to obtain exact values of peak current. This results from the fact that the sine wave veers away from a linear rise, the departure making significant differences for greater melting angles than 30 degrees. A detailed treatment of this subject is contained in part II of this paper.[1]

The peak current at melting is one of the most important factors contributing to test severity. In view of this alone, it would be expected that higher available short-circuit current leads to greater severity in interruption. Existing standards[13] state that tests to establish time–current characteristics should be made at room temperature with no initial load current. Similarly, interruption tests should be made with no preloading, which would cause a reduction in the peak current at melting and in the melting time. However other factors contribute to the severity, notably, the magnitude of the circuit voltage during the early arcing period. It is during this time that a current-limiting power fuse either succeeds or fails to develop sufficient arc resistance, hence arc voltage drop, to exceed the generated voltage and cause the current to be brought back to zero. The arc voltage must exceed generated voltage during this critical interval in order to effect a reduction of arc current and subsequently create the current zero necessary for interruption. It is possible for arc voltage to be less than generated voltage for short intervals during the arcing period and still obtain interruption. The arc current increases during such an interval, as in the borderline interruption of Figure 4B. Test records of this type are invaluable to the fuse designer, as they indicate the threshold of safe operation and the direction in which design features must be changed to provide the desired factors of safety.

The energy released in the fuse during the arcing period is another important severity factor and may be computed by integrating the product of arc voltage and arc current over the arcing period. Arc energy is not necessarily a maximum at the maximum available short-circuit current, and for this reason it is recommended that interruption tests be made not only at the current representing the interrupting rating of the fuse unit, but at several intermediate values to cover any variations in the energy released in the fuse unit itself. A full discussion of the effect of circuit variables upon the energy absorbed by the fuse is included in part II of this paper.[1]

Low Current Operation

The ability to interrupt low available currents barely capable of causing fuse operation, is expected of power fuses manufactured to meet existing standards.[13] Depending on its heat capacity, a fuse may require between a few minutes and an hour to reach its final temperature and have its element melted.

Existing National Electrical Manufacturers Association Standards, call for the assignment of ampere ratings on the basis shown in Table I.

A sufficient number of interruption tests should be made in the range of such low available currents to prove the ability of the fuse unit to interrupt under these conditions. The most likely causes for prolonged overload currents of such magnitudes are internal transformer faults, usually in the insulation between turns of the primary winding, leading to a progressive drop in primary impedance. It is estimated that the power factor of such

Table I

Ampere Ratings	Per Cent of Rated Current	Melting Time, Minutes
Up to and including 100 *E* amperes*	200–240	5
125 *E* amperes* and above	220–264	10

*Current ratings include the letter *E* to denote conformance with the standards.

faults will be at least 20 per cent; hence it is recommended that the test circuit have a power factor of 20 per cent maximum.

Fault Starting Angle

The angle of the generated voltage, at the instant of inception of the short circuit, has an important influence on severity. Until melting occurs, the only effect of this fault starting angle is to produce displacement of the available current. If a highly inductive circuit is closed at an instant when voltage is zero, the displacement will be the maximum, and the available current will be fully asymmetrical (Figure 5). With the fault starting at 90 degrees (peak voltage), symmetrical available current will result. For intermediate starting angles, the displacement will be partial, and a "major loop" of available current will be obtained. A starting angle greater than 90 degrees (or after peak voltage) will result in a "minor loop" of current which affects the fuse as a symmetrical current of smaller magnitude and higher frequency.

In Figure 5 the degrees of asymmetry are compared for different fault starting angles. The available short-circuit current is 20,000 symmetrical rms amperes, and peak currents obtained are indicated for typical current limiting fuses of various ampere ratings.

For a fuse of low ampere rating, maximum peak current is obtained on a symmetrical short circuit since its initial rate of rise is greater than could be obtained with any asymmetry. Although the symmetrical current has the greatest initial rate of rise, the melting angle of fuses of large ampere ratings can be so great that the departure from a linear rise becomes significant, this effect taking place over the same range of voltage angles at which the asymmetrical current is approaching its maximum slope. Thus, if the ampere rating of the fuse is large enough, the peak current obtainable on an asymmetrical current can be even greater than that obtained on a symmetrical current. In the case of the 200-ampere fuse unit (Figure 5) melted on a circuit capable of producing 20,000 symmetrical rms amperes, slightly higher peak currents can be obtained when the current is asymmetrical than when it is symmetrical. It is also interesting to note that in the range between, in which major loops of short-circuit currents are obtained, the currents can be slightly higher than either of these values because the effective slope is higher during the melting interval.

Melting Angle

It is to be expected, and it has been borne out by tests, that maximum severity occurs in fuse units of low ampere rating if melting takes place at, or slightly before, peak voltage. For such a fuse, this is also the closing angle producing maximum peak currents. This would call for closing the circuit at approximately 80 degrees for such ratings. The higher ampere ratings require a considerable number of degrees to melt. An example of this is the case of the 200-ampere fuse in Figure 5 which requires approximately 76 degrees to melt on the 20,000-ampere asymmetrical short circuit. It has been found empirically and it is shown analytically in part II of this paper,[1] that maximum severity is obtained if the circuit is closed at such a voltage angle as to cause melting to take place in the region between 70 and 80 degrees. This results in high effective generated voltage during the critical early arcing period.

In Figure 6 oscillograms are shown of three tests on the same circuit, on three developmental fuse samples of high current rating, the circuit being closed at different angles after voltage zero. The test of Figure 6A was evidently the most difficult, as the short-circuit current was not interrupted, and the fuse sample broke apart as a result of the high release of energy inside the fuse tube. In this test, the circuit was closed at 11 degrees and melting took place at 68 degrees. At this time, circuit voltage was approximately 90 per cent of maximum and still increasing. The subsequent attempt on the part of the fuse to produce arc voltage equal to effectively maximum circuit voltage was unsuccessful. In Figure 6B the circuit was closed at 90 degrees of voltage producing a symmetrical current, but melting did not occur until 138 degrees, when voltage was already down to 67 per cent of maximum voltage. In Figure 6C the oscillogram shows a relatively easy interruption resulting from closing the circuit after voltage maximum, resulting in a minor loop of current, with melting occurring in the following half cycle. Although circuit voltage was near maximum when melting did take place, the melting current was somewhat less in this test and alleviated the severity. None of the possible fault starting angles after 90 degrees have been found to cause as much test severity as those before 90 degrees.

Circuit Frequency

The frequency of the test circuit determines the time interval between voltage zeros and therefore has an important bearing on test severity. This leads to the recommendation that interruption tests be made at rated frequency. In the time between melting and the subsequent voltage zero, a current limiting power fuse must develop enough arc voltage drop to force the current down to zero before excessive energy is released within the fuse tube.

On a circuit of lower frequency, the longer time between voltage zeros brings with it an increase in test severity, other factors being equal. This is reflected in

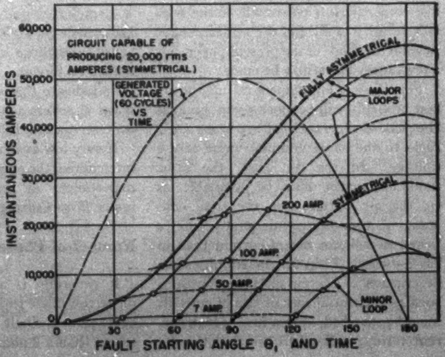


Figure 5. Effect of fault starting angle on displacement of available current

Melting peak current of typical current limiting power fuses of 7-, 50-, 100-, and 200-ampere ratings

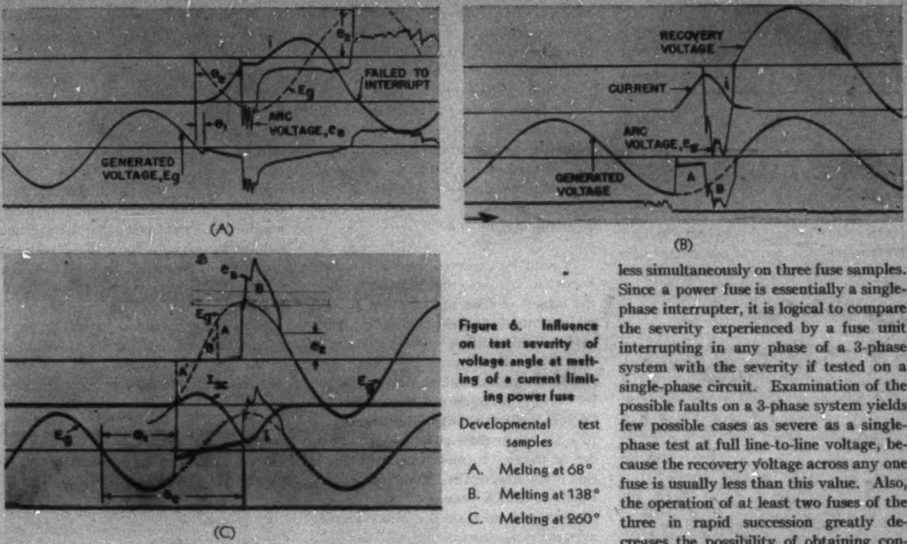


Figure 6. Influence on test severity of voltage angle at melting of a current limiting power fuse

Developmental test samples

A. Melting at 68°

B. Melting at 138°

C. Melting at 260°

lower interrupting ratings at 25 cycles than at 60 cycles for many present-day high current rated fuses. Figure 7 shows oscillograms of tests made at 25 and 60 cycles at the same circuit voltage, on the same design of a high current rated fuse test sample. In all other respects except frequency, the 60-cycle test would be regarded as more severe because the available and melting currents were higher, and because melting occurred nearer maximum voltage. However, the lower frequency test was clearly more severe, as evidenced by the higher arc current (Figure 7A) following melting and the longer arc duration.

A d-c circuit is the extreme case of low frequency. Here the fuse must force a current zero without any help from voltage zeros. A full treatment of the relative severity of a-c and d-c test circuits is beyond the scope of the present paper, but a few observations may be made.

For a fuse of low ampere rating, the a-c test probably will be more severe because the peak of generated voltage bears a $\sqrt{2}$ ratio to the same nominal value of d-c voltage. In the d-c case, the same nominal current may be available as in the a-c circuit, but the time constant of the d-c circuit may cause a faster or a slower rise of current than in the a-c case. Thus for d-c circuits some description of the circuit time constant, such as the initial rate of rise of current, must be specified. For a fuse of high current rating, indications are that the d-c test might be more severe on account of the absence of voltage zeros.

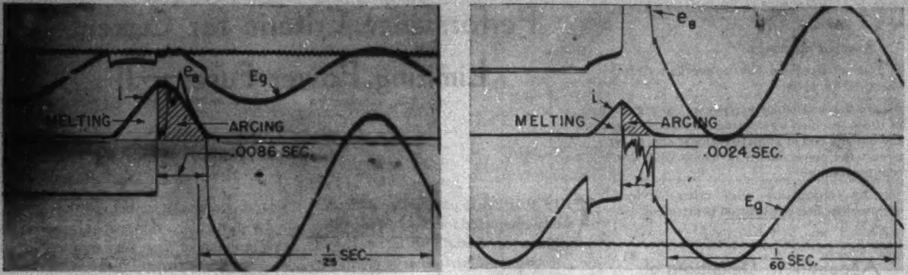
Power Factor

To some extent the power factor of the test circuit will have an effect on severity. At a low power factor with a symmetrical short-circuit current, a fuse of low ampere rating which melts quickly will have nearly a 90-degree wait until the next voltage zero. At higher power factors, with the current and voltage more nearly in phase, the delay of the subsequent voltage zero is even longer, up to the maximum of 180 degrees. For a fuse of high current rating, severe test conditions apparently will result if the power factor is such that voltage is in the critical 70- to 80-degree region at melting. Tests have been made at high power factors by inserting resistance in an otherwise highly inductive circuit. The results of such tests indicate that the interrupting duty is not as severe as that of low power factor tests made at full short-circuit current.

If very low available currents are to be interrupted, the power factor has a greater importance, as mentioned previously under "Low Current Operation."

Number of Phases

It generally is agreed that a single-phase test made with full line-to-line voltage across a single fuse sample is more severe than a 3-phase test, made more or less simultaneously on three fuse samples. Since a power fuse is essentially a single-phase interrupter, it is logical to compare the severity experienced by a fuse unit interrupting in any phase of a 3-phase system with the severity if tested on a single-phase circuit. Examination of the possible faults on a 3-phase system yields few possible cases as severe as a single-phase test at full line-to-line voltage, because the recovery voltage across any one fuse is usually less than this value. Also, the operation of at least two fuses of the three in rapid succession greatly decreases the possibility of obtaining conditions of maximum severity on either fuse. In Figure 8A oscillograms are shown of a 3-phase interruption of a triple line-to-ground fault (generator windings connected in delta with no ground connections) by current-limiting power fuses, and Figure 8B is the record of a single developmental sample fuse of the same design tested line-to-line on the same circuit. The single-phase test was more severe, because it resulted in failure, and because the records of current in each phase show (Figure 8A) definite current-limiting action. In the 3-phase test there were no signs of high arc current, a characteristic of difficulty in interrupting. Although the short circuit was closed with a 3-phase circuit breaker, not all the poles closed exactly at the same time. Even with simultaneous closure of the three lines at the fault, unbalanced currents generally would result because of phase differences in the generated voltages. In the 3-phase test of Figure 8, the fuse in phase one operated first and opened phase one before the fuse in phase two melted, and at approximately the same time as the fuse in phase three melted. This left the fuses in phases two and three effectively in series, and they then assisted each other in clearing the circuit, the burden being apparently more on the fuse in phase three, as may be noted from the small rise in arc current after melting, and the corresponding dips in arc voltage in phase three. Phase two apparently cleared easily as may be noted

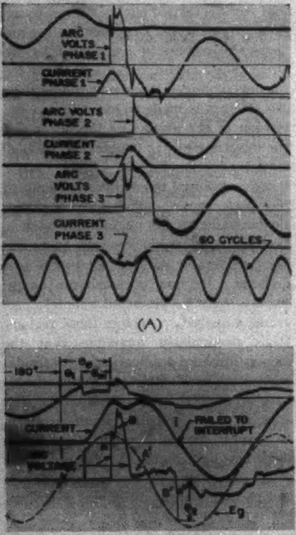


(A)

(B)

Figure 7. Effect of circuit frequency on test severity

	A	B
Circuit frequency........	25 cycles..	60 cycle
Test voltage (rms)........	2,500 volts ..	2,500 volts
Available current (rms)...	20,000 amp ..	25,000 amp
Current at melting........	18,000 amp ..	28,000 amp
Angle of inception......	62° ..	50°
Angle at melting........	136° ..	108°
Arcing time..............	0.0086 sec ..	0.0024 sec



(A)

(B)

Figure 8. Single-phase and 3-phase interruption tests on the same current-limiting power fuse design showing the greater severity of the single-phase test

	A 3-Phase	B Single-Phase
Frequency, cycles....	25	25
Test voltage (rms), volts..............	2,500......	2,500
Available current (rms) including displacement, amperes	Phase 1—39,000 Phase 2—30,000..... Phase 3—25,000	28,000
Peak amperes at melting (instantaneous)	Phase 1—21,000 Phase 2—12,200..... Phase 3—13,400	19,800
Series reactance, ohms..............	Phase 1, 2, 3, 0.04....	0.04
Performance......	Fuses in all three phases interrupted without disturbance	Fuse failed to interrupt the circuit

from the concave shape of the arc current and the resultant arc volts which remained well above generated voltage from melting onward. This mutual assistance by the phases is characteristic of 3-phase tests, and in many similar tests has resulted in the ability to interrupt a greater nominal kilovolt-amperes than the 3-phase kilovolt-amperes computed from single-phase interrupting ability.

The symmetrical rms short-circuit current obtainable by making single-phase tests phase-to-phase on a 3-phase generator is only 86 per cent of the line current obtainable in a 3-phase short circuit. It is recommended that the 3-phase kilovolt-ampere interrupting ability be computed from the measured current interrupted by a single fuse tested line-to-line. It is believed that this basis of rating will cover all short-circuit conditions (whether faults are single line to ground or phase-to-phase) and justify the use of the maximum rms ampere rating, including cases where single-phase applications are proposed, and the single-phase interrupting requirements are known.

Specific Applications

In some cases, a given rating of power fuse is designed for a specific application rather than being made available for general use. Typical cases are

1. Fuses for high voltage motor controls. For such cases, not only the interrupting capacity of the fuse unit itself must be determined, but tests must be made with the contactor and current transformer in the circuit to check the ability of these devices to windstand the let-through current of the fuse, both from a thermal and magnetic standpoint. The phrase let-through current implies not only the peak current permitted by the current limiting action of the fuse, but also the magnitude and duration of arc current until interruption is completed.

2. Fuses for use with banks of capacitors. For these, safe limits of application must be determined, within which the fuse let-through current will not cause the case of a faulted capacitor unit to be ruptured, endangering apparatus and personnel, and possibly allowing the fault to become external and spread to the other phases.

Conclusions

In establishing the interrupting ability of current-limiting power fuses, it is recommended that tests be made as follows:

1. In a short-circuit testing laboratory where circuit variables can be controlled and measured. Tests on systems may give additional data, providing adequate control and measuring means are used.

2. Single-phase.

3. At the fuse unit rated voltage in accordance with established standards.

4. With the starting angle of the short-circuit current at zero of voltage to justify the maximum rms ampere interrupting rating (asymmetrical—including d-c component) to be assigned.

5. With the starting angle of the short-circuit current near 90 degrees after voltage zero (preferably earlier) in the case of fuse units of low ampere ratings, and at such a current as to cause melting of the fusible element between 70 degrees and 80 degrees on the voltage wave, for fuses of high ampere ratings.

6. At several other values of short-circuit current less than the interrupting rating, the angle of inception being in each case such as to cause melting somewhat before maximum voltage. Three-phase kilovolt-ampere interrupting ability, when assigned, to be calculated from the symmetrical a-c component of short-circuit current I, successfully interrupted single-phase as in pre-

ceding 4 and 5, that is, kilovolt-amperes = $\sqrt{3}\times$ kilovolts $\times I$.

7. At rated frequency.

8. On a circuit of low power factor, for short-circuit tests.

9. At low available currents in the region of those specified by existing standards[13] for the assignment of ampere ratings, and on a circuit the power factor of which is 20 per cent maximum.

10. In conjunction with other devices with which their application is intended.

Appendix I. Definitions of Symbols Shown on Oscillograms

E_g = generated voltage
e_a = arc voltage of fuse
I = available symmetrical rms short-circuit current
I_{sc} = short-circuit current
i = actual circuit current
θ_1 = fault starting angle
θ_a = arc voltage angle
$\theta_m = \theta_a - \theta_1$ = melting time of fuse in degrees
θ_3 = angle when arc current becomes zero (interruption)
e_3 = instantaneous generated voltage at θ_3
A, B, A', B' = volt–time areas

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Performance Criteria for Current-Limiting Power Fuses—II*

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THIS PAPER is part of an analytical study conducted for the purpose of establishing a sound fundamental interpretation of observed performance characteristics of high interrupting-capacity power fuses of the current-limiting type. Some of the more important conclusions, which have been verified by a study of test performance, are presented and discussed in the foregoing part of this paper by C. L. Schuck. In another sense this paper is an application of the analytical methods developed in the two papers on "Geometry of Arc Interruption."[15,16] This application to fuse performance is particularly fitting since the interval from the inception of a short circuit until the arc voltage of the fuse is introduced, may be predetermined quite accurately from the melting characteristics of the fuse. Conductor heating and energy equations are presented for the first time which include all the essential variables to predetermine the arc energy to be developed by the interrupter for various magnitudes of arc voltage, degrees of circuit asymmetry, and melting times of the fuse.

Although the analytical methods used in this paper are applied to the analysis of current-limiting power fuse performance, they are by no means restricted to this application. When applied to a circuit breaker, for example, the so-called melting time of the fuse (Θ_m) becomes the contact parting time of the circuit breaker. Indeed, many high speed interrupters[17] whose contact parting times may be predetermined accurately are subject, in some degree, to the analysis and conclusions here presented. One such application of this treatment to power circuit breakers for rectifier service has been presented previously[17] before the AIEE.

In particular, this paper approaches the problem of fuse performance on an analytical basis, to establish certain fundamental trends which result as one or more of the circuit variables is altered. The work is carried out on a quantitative basis in such a way as to enable the equations and curves to be applied to any single-phase circuit regardless of the frequency, voltage, or current magnitudes. No attempt is made to enter into a discussion of the physics of the arc within the interrupter or to explain the methods used to develop the arc voltage characteristics essential to good current-limiting fuse performance. The study is of particular value to the testing and application of the interrupter as well as rendering considerable data of use to the fuse designer. Regions of maximum circuit severity may be predicted and test programs arranged to explore these regions. It is particularly fortunate that the straight forward application of the fundamental concepts of linear circuit theory may be applied to solve what otherwise appears as a nonlinear circuit problem. In several places this analytical study sheds new light on the performance characteristics of current-limiting power fuses and clears up what otherwise might ap-

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*Part I of this paper was prepared by C. L. Schuck.

pear to be inconsistencies in the laboratory and field performance of this interrupter.

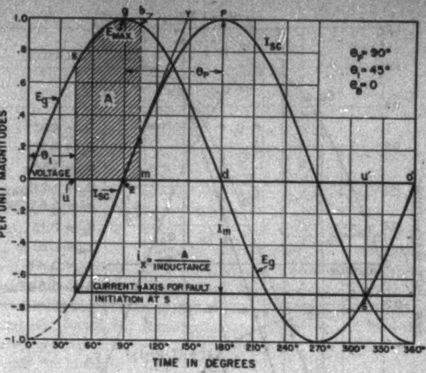


Figure 1. Relationship between volt-time areas and short-circuit current in a single-phase zero-power-factor a-c circuit

The short-circuit current i_x is equal at all times to the volt-time area under the generated voltage. Note that the current axis depends upon the angle Θ_1

Assumptions and Nomenclature

Any quantitative analytical treatment which is devoid of empirical data reincorporated into the analysis is restricted by certain assumptions which usually are made to facilitate the treatment. In spite of these assumptions, however, it is the duty of the analyst to render a solution which indicates the salient trends to be found in the device under study, although it is recognized that no one single performance ever will be duplicated exactly in the analytical study. In the study under consideration, it will be shown, for example, that the ratio of the magnitude of the arc voltage developed by the interrupter to the generated voltage is a controlling factor in its performance. For this reason the study includes a solution for numerous assumed ratios of arc voltage to generated voltage and such results are plotted with this ratio as the variable.

Because the study at hand is concerned with fuses applied for the interruption of short-circuit currents, the assumption of a zero power factor circuit is made for the great majority of the data developed. It is to be made clear, however, that the analytical method employed is not limited to the zero power factor circuit. Moreover, it is not to be implied that zero power factor always presents the most severe condition in all cases.

Field, laboratory, and analytical studies all confirm that the fuse experiences its most difficult tasks when clearing single-phase short circuits. For this reason (as discussed more fully in the foregoing part I of this paper by C. L. Schuck), the analytical treatment presented here is concerned entirely with single-phase interruptions.

In order that the relations developed might be applied to circuits of any voltage, any current, and any frequency, a per unit system of nomenclature was selected which expresses the arc energy as a coefficient k which must be multiplied by the circuit MVA (based upon a symmetrical current) to obtain the true arc energy in watt hours. Similarly, the time scales are expressed in degrees in order that the results might be applied to circuits of any frequency. The reference point for measuring the major angles, namely, Θ_1, Θ_e and Θ_2, as listed following, is the zero of the generated voltage. This has many advantages in the mathematical treatment. The salient nomenclatures as used in the paper are

Θ_p = power factor angle
  90 degrees = all inductive, no resistance
Θ_1 = fault starting angle (measured from voltage zero)
  90 degrees = symmetrical fault current when $\Theta_p = 90$ degrees
  0 degrees = fully displaced current
Θ_e = arc voltage angle (measured from voltage zero)
$\Theta_m = (\Theta_e - \Theta_1)$ = melting time of fuse
Θ_2 = angle when final current zero occurs (measured from voltage zero)
$\Theta_B = (\Theta_2 - \Theta_e)$ = arcing time
$\Theta_f = (\Theta_2 - \Theta_1)$ = fault duration
E_g = generated voltage (single-phase, rms)
E_m = crest of generated voltage
e_B = arc voltage of interrupter in per unit of E_m
e_2 = instantaneous generated voltage at interruption in per unit of E_m
I = available symmetrical, rms, short-circuit current (no arc voltage)
I_{sc} = short-circuit current available (includes d-c component)
I_m = peak value of available current
i_B = current that would flow if e_B acted alone in the circuit
i = actual current in the circuit $= I_{sc} - i_B$
i_m = peak value of actual current
Q = the integral of arc current over the arcing time
L = inductance of the circuit in henries
$MVA = E_g \times I \times 10^{-6}$ = available, symmetrical, zero-power-factor, megavolt-amperes
k = arc energy coefficient, watt-hours per megavolt-amperes (60-cycle circuit)
k_L = maximum inductive energy in circuit in watt-hours per megavolt-amperes
W = arc wattage in watt-hours $= k \times$ megavolt amperes (60-cycle circuit)
$\omega = 2\pi f$
f = circuit frequency
$\Sigma i^2 t$ = summation $i^2 t$ of a fuse. This is obtained from the basic fuse design data and concerns the material and cross section of the fusible element. See equation 7
$H = \dfrac{\Sigma i^2 t \times \omega}{I^2}$ = heating function (equation 5)
x_L = inductive ohms of circuit
  $= \omega L = E_g / I$
A, A', B, B', C = voltage–time areas

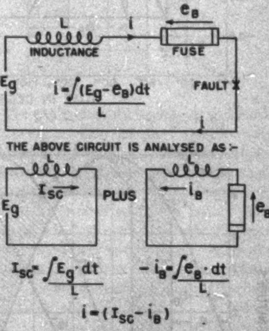

e_B
L
i
INDUCTANCE
FUSE
E_g
$i = \frac{\int (E_g - e_B)\,dt}{L}$
FAULT
THE ABOVE CIRCUIT IS ANALYSED AS :-
L
I_{SC}
PLUS
L
i_B
e_B
E_g
$I_{SC} = \frac{\int E_g \cdot dt}{L}$
$-i_B = \frac{\int e_B \cdot dt}{L}$
$i = (I_{SC} - i_B)$

Figure 2. Fundamental circuits used in the analytical determinations in this paper

Here the principle of superposition is used to show the concept of a backward moving or cancellation current produced by the fuse voltage

Analytical Approach

The student of short-circuit protection will achieve a clear insight into the behavior of interrupting devices by applying a few fundamental laws of circuit theory. The simplest of these concerns the voltage-current relationship in a purely inductive circuit. The majority of a-c short circuits is limited primarily by the inductive reactance of the circuit. Short-circuit current calculations on transmission systems usually neglect circuit re-

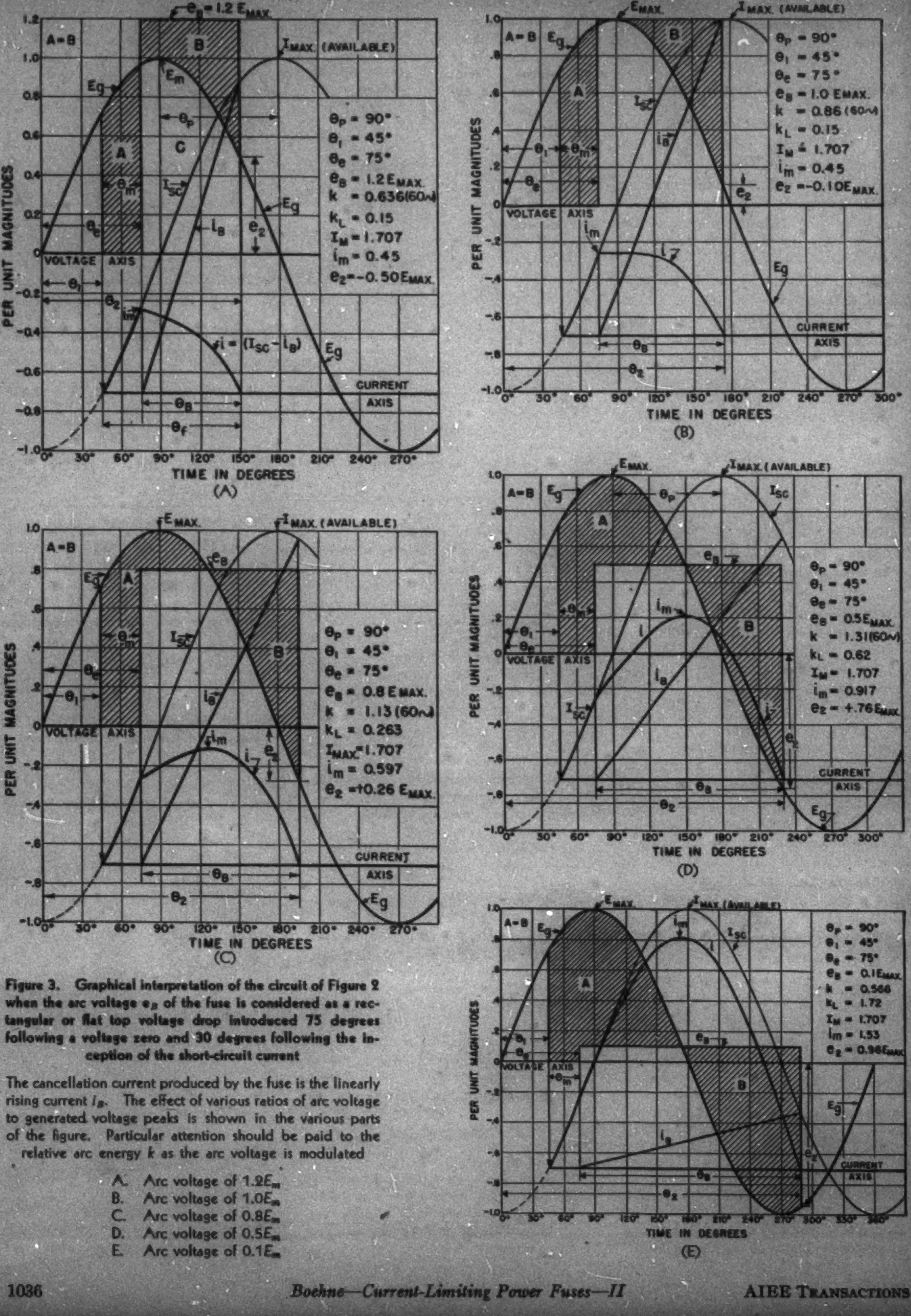


Figure 3. Graphical interpretation of the circuit of Figure 2 when the arc voltage e_B of the fuse is considered as a rectangular or flat top voltage drop introduced 75 degrees following a voltage zero and 30 degrees following the inception of the short-circuit current

The cancellation current produced by the fuse is the linearly rising current I_B. The effect of various ratios of arc voltage to generated voltage peaks is shown in the various parts of the figure. Particular attention should be paid to the relative arc energy *k* as the arc voltage is modulated

A. Arc voltage of $1.2E_m$
B. Arc voltage of $1.0E_m$
C. Arc voltage of $0.8E_m$
D. Arc voltage of $0.5E_m$
E. Arc voltage of $0.1E_m$

sistance and can do so with a high degree of accuracy. Although the most pessimistic conditions cannot be predicted always on the basis of the purely reactive circuit (zero power factor), nevertheless a working understanding of the purely inductive circuit forms the foundation of this phase of the protection engineers' art. Just as the current in a noninductive circuit is given by E/R so the current in the purely inductive circuit may be expressed as the voltage-time area divided by the circuit inductance. The fundamental equation of current in the purely inductive circuit hence becomes

$$i=\frac{\int e\times dt}{L} \qquad (1)$$

This is just another way of expressing the well-known voltage drop in an inductance, namely,

$$e=L\frac{di}{dt}$$

This volt–time area concept is only applicable to the purely inductive circuit and is the foundation of a simple geometry which requires that from current zero to current zero the net area under the circuit voltages is zero. Figure 1 shows this law applied to the generation of the reactive short circuit in a purely inductive circuit, giving a pictorial understanding of the cause of the asymmetrical short circuit. The voltage E_g in Figure 1 is the fixed sinusoidal a-c generated voltage with its fixed axis. The current I_{sc} is the short-circuit current, the current axis being determined by the instant of short-circuit initiation. For example, if the fault starts at s, on the generated voltage, a vertical from the voltage to the current determines the current axis for that condition. The magnitude of the current at any instant is then proportional to the volt–time area from fault inception up to that instant. The current i_s, for example, is equal to the volt–time area A = s-b-m-u-s divided by the circuit inductance as defined in equation 2. It follows that in this case, peak current I_m will occur at a voltage zero, its magnitude being dependent upon the volt–time area s-d-u. Hence current zero will occur when an equal area under the negative voltage has been covered, that is, when area $s'du'$ = s-d-u. The magnitude of the minor loop of current which follows will be proportional to the area $s'o'u'$. Inspection will reveal now that a fault starting at voltage peak will produce a symmetrical fault current (constant inductance) having a peak value d-P. Similarly a fault starting at a voltage zero will produce a fully displaced short-circuit current having a peak twice this magnitude. These area concepts therefore produce results which agree with other approaches to the same problem and will be extended now in simple graphical form to cover the performance of fuses.

When more than one voltage acts within a circuit, the current resulting from each source can be determined independently and the resulting currents added to determine the net current. A study of Figure 2 reveals this principle applied to the problem at hand. This superposition of currents is well known and often applied to the solution of circuits in cases when the multiple source of voltages are all generated electromotive forces and batteries. The broadening of the superposition principle to consider voltage drops acting in the circuit just as if they were generated electromotive forces generally is not used and its power is not appreciated fully. This extension of the fundamental superposition principle, as an analytical tool, finds its greatest usefulness in the analysis of many circuits containing devices whose voltage drops are nonlinear, such as tubes, circuit interrupters, arc furnaces and welders. This is particularly true when the performance of these devices depends upon their voltage drops and this characteristic, in turn, can be controlled by the designer. Such is the case with most circuit interrupters and in particular with power fuses of the current-limiting type.

The voltage drop characteristics of a power fuse determine the energy liberated in the fuse, the recovery voltage which appears following interruption, its current-limiting ability and, in general, controls the performance of the fuse. These facts may be demonstrated most easily by considering the voltage across the fuse to take a rectangular shape following the melting of the current-responsive element in a zero power factor circuit. The current i_B (Figure 2) produced by a rectangular or d-c voltage, e_B, in a purely inductive circuit is, as outlined before, proportional to the area under the voltage. This current, i_B, is hence a linearly rising current, at a rate e_B/L, and given in this case by the equation

$$-i_B=\frac{\int e_B dt}{L}$$

$$=\frac{e_B\times t}{L}$$

The scale of this current, for graphical analysis, is obtained most readily from Figure 1 by observing that, for a symmetrical fault originating at crest voltage, E_m = xg, a linearly rising current xy would have resulted in one radian (57.3 degrees) had the crest voltage E_m been maintained over the interval. When the fuse voltage, e_B, is rectangular, it produces a current which rises at a rate e_B/E_m times the slope of xy on Figure 1. It follows

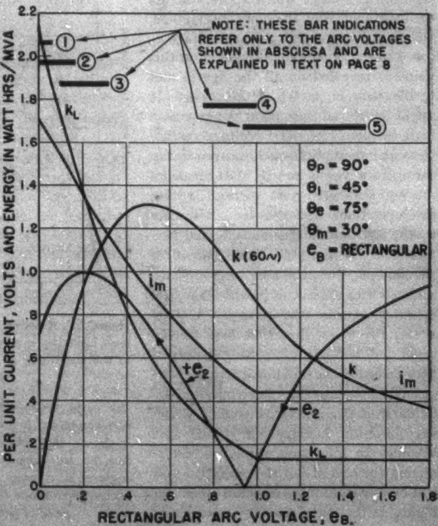


Figure 4. Master plot of the essential data from Figure 3 as a function of the magnitude of the ratio of arc voltage to generated voltage crest

Note that peak arc energy occurs for an arc voltage whose magnitude is about 0.5 of the crest of the generated voltage. The horizontal bar lines at the top of the figure pertain only to the nominal relative arc voltage range of the following interrupters:

1. Air blast breaker
2. Oil circuit breakers
3. Expulsion fuses
4. Magne - blast breakers
5. Current - limiting power fuses

that the current i_B, so developed as a result of the fuse voltage e_B, is a current which flows in opposition to the short-circuit current I_{sc} (see Figure 2) and is plotted in all figures in a positive sense to visualize better its algebraic addition to I_{sc}. A typical example of such an analysis is shown in Figure 3A which will be described in detail as a demonstration of the method used to determine all the analytical data developed in this paper. Here a fault is assumed to originate in a zero power factor circuit ($\Theta_p = 90$ degrees) at an interval of $\Theta_1 = 45$ degrees following the voltage zero. This would have produced a short-circuit current, I_{sc} rising to 1.707 times the normal symmetrical current crest. At an interval of 30 degrees following current initiation, the current-responsive element in the fuse melts ($\Theta_m = 30$ degrees) and gives rise to an assumed rectangular arc voltage e_B of 1.2 times the generated voltage peak. This arc voltage accordingly would produce the linearly rising cancellation current i_B which rises at a slope of 1.2 times the slope of the short-circuit current at maximum voltage (1.2 sy on Figure 1). When this current i_B becomes equal to the short-circuit current I_{sc}, that is where the two currents intersect on the diagrams in this paper, current zero occurs. The fuse current i, following the melting of the current-responsive element, is therefore the difference between the two currents I_{sc} and i_B. This resulting current is shown plotted on the current axis. As current zero is reached, and if interruption occurs, the voltage which appears across the fuse is naturally the generated voltage. Hence at interruption the arc voltage e_B drops to the generated voltages E_g at this point as revealed on all the figures and oscillograms in part I of the paper. It might be advisable, to point out here that high frequency recovery voltage oscillations at interruption, such as accompany the operation of oil and air blast breakers, are not associated with current-limiting fuse operation. This is borne out further by the fact that current-limiting fuse performance is essentially independent of the amount of capacitance to ground associated with the circuit in which the fuse operates. Hence the recovery voltage across the fuses, discussed here, is the generated voltage. The fuse, however, does recognize the initial magnitude of the generated voltage e_2 at interruption, particularly if the generated voltage is rising following interruption. This fact becomes an important element in determining the severity of the operation and will be discussed more in detail later. In Figure 3A the initial recovery voltage e_2 is $-0.50E_m$, the minus sign indicating that the current zero leads the voltage zero or that the generated voltage is falling. This represents a favorable operation.

Area Criterion

Since current is proportional to the volt–time area covered in the purely inductive circuit, it follows that equal volt–time areas must have been covered by both the generated voltage E_g and the arc voltage e_B at the instant the two currents become equal. The volt–time area $A+C$ must be equal, therefore, to the volt–time area $C+B$ in Figure 3A, and accordingly $A=B$. This fact finds mathematical expression in the equations found in Figure 2. From these facts emerge some of the more obvious general conclusions as can be deduced by the reader by an inspection of Figure 3A, namely,

1. The arc voltage of the fuse must exceed the generated voltage and remain above the generated voltage for a time sufficiently long to create a volt–time area B equal to the volt–time area A developed during the melting time of the fuse.

2. Fuses of low current rating and thus having small melting times (θ_m) will produce a small volt–time area A and accordingly make it easier to create the neutralizing area B resulting from fusion.

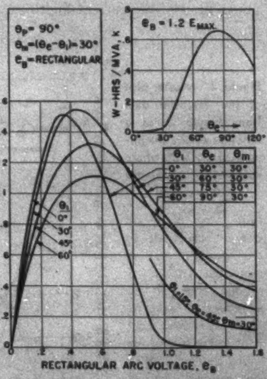


Figure 5. Relative arc energies for various fault initiating angles Θ_1 plotted with respect to the magnitude of a rectangular arc voltage ratio e_B

The curve for $\Theta_1 = 45$ degrees is replotted from Figure 4. In the upper right-hand corner is shown the relation between arc energy and the arc voltage angle Θ_e for a melting angle of 30 degrees. Note that peak energy occurs when fusion takes place just before the peak of generated voltage

3. Because the arc voltage is controlled by the fuse design, it follows that fuses applied in circuits having higher than the rated voltage of the fuse will make it difficult to create the area B and in any event the current zero will be delayed.

These facts will be made clearer and some of the less obvious facts will emerge with the modulation of some of the variables associated with fuse performance.

Arc Energy

The energy released by a current-limiting fuse preferably is absorbed within the fuse without any display of expelled gases or mechanical breakage. This is one of the reasons why the energy released is important to the fuse designer. Of more importance, however, is the influence of the energy released upon the interrupting ability of the fuse. Arc voltages of sufficient magnitude are maintained by cooling the arc path at a rate faster than the energy generation. Experience shows that the more energy released per conductor, the lower the arc voltage becomes; and, as will be shown, the lower the arc voltage, the more energy is released. The fuse designer, therefore, must proportion the materials so as not to produce too high a voltage surge at fusion, yet produce as nearly as possible the optimum arc voltage characteristic for over-all fuse performance. This is done by the proper choice of multiple conductors, their materials, diameter, length, and arrangement, as well as the material and grain size of the fuse filler.

Returning now to Figure 3A, the arc wattage is the product of arc voltage e_B and the arc current i. Because the arc voltage is constant in this case, the wattage is proportional to the arc current i and hence is not replotted. The arc energy is proportional to the watt–time area, that is, the area under the wattage curve. For those cases where the arc voltage is constant during the arcing period, the arc energy liberated by the interrupter in a 60-cycle zero power factor circuit, is given by

$$k = \left[1.4736 \times e_B \times \int_{\Theta_e}^{\Theta_2} i\,dt\right] \text{ watt-hours per megavolt-ampere per half cycle of arc.}$$

(see equation 13)

This integral of arc current and time is termed Q and is shown in Appendix I to be the following function of the angles of the circuit

$$Q = \int_{\Theta_e}^{\Theta_2} i\,dt$$

$$= {}^1/_2\Theta_B(\cos\Theta_1 + \cos\Theta_2) - (\sin\Theta_2 - \sin\Theta_e) \qquad (2)$$

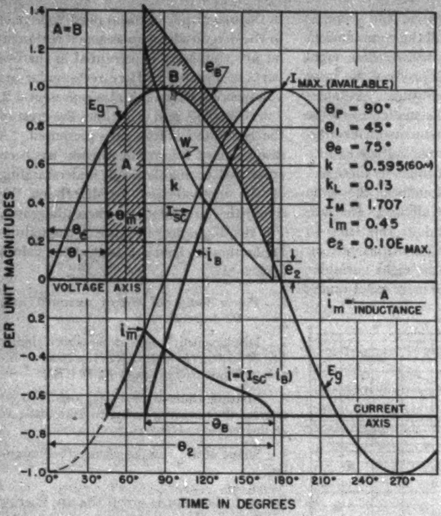


Figure 6. Although the arc voltage e_B differs from a rectangular shape, the area criterion ($A=B$) must hold from current zero to current zero in a zero-power-factor circuit

Here the interrupting current zero was chosen to take place at the same time as found in Figure 3B. Note that the arc energy above is less than in Figure 3B, all other factors being identical save the arc voltage wave shape. Figure 7 shows what happens, however, when such a characteristic is used in a 25-cycle circuit. See Figure 6C in part I for oscillogram showing these characteristics

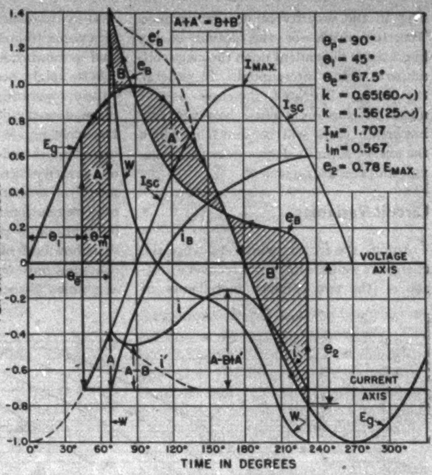


Figure 7. When the fuse characteristic of Figure 6 functions in a 25-cycle circuit, the melting angle is only 22.5 degrees

The drooping arc voltage wave shape however causes the arc voltage to dip below the generated voltage as indicated by the area A'. During the interval A', therefore, the arc current must rise and only come to zero as the area B' is created such that $A+A'=B+B'$. Note that the arc energy k and the initial recovery voltage e_2 both represent more severe conditions than the corresponding 60-cycle values on Figure 6. See Figure 8B in part I for oscillogram showing these characteristics on a 25-cycle circuit

A more general expression of arc energy which applies for any frequency is given in Appendix I, equation 12, and repeated here

$$\text{arc energy} = 555\frac{e_B Q}{\omega} \text{ watt-hours per megavolt-ampere} \quad (12)$$

The conditions for which these arc energy equations are valid are indicated clearly in Appendix I.

The use of these equations requires the determination of the angle Θ_2, or the terminating angle of the current. An inspection of Figure 3 will show that this angle is related to the arc voltage e_B by the following relationship

$$e_B = \left(\frac{\cos\Theta_1 - \cos\Theta_2}{\Theta_B}\right) \quad (3)$$

Because $\Theta_B = (\Theta_2 - \Theta_1)$ it follows that no exact solution for Θ_2 is possible because of the transcendental nature of the equation. Graphical methods, as disclosed in Figure 3, will give a very close approximation to Θ_2. A more exact determination can be achieved quickly with one or two insertions in the preceding equation. Mass calculations assuming Θ_2 as the variable and later cross plotting for e_B are quite useful. The preceding arc energy relations are plotted in Figure 9 as a function of the melting time of the fuse for a particular arc voltage e_B of $1.2E_m$ in a 60-cycle circuit. The actual energy W developed by the fuse is accordingly

$$W = k \times MVA \text{ watt-hours}$$

The arc energy represented by Figure 3A is 0.636 watt-hour per megavolt-ampere. Hence a fuse operating on the pattern of Figure 3A in a 60-cycle circuit interrupting 36,000 amperes at 2,800 volts would develop 64 watt-hours. The peak energy stored in the circuit inductance during this operation is $1/2Li_m^2$ and is termed the maximum inductive energy k_L. This is given on Figure 3A as 0.15 watt-hour per megavolt-ampere. Hence the energy developed within the fuse exceeded the stored energy by 425 per cent.

The preceding energy values come into proper perspective with a few limiting values indicated. Consider a single-phase zero power factor circuit, representing a definite short-circuit megavolt-ampere, and allow a linear resistance of ohmic value equal to the inductive ohms to be added in series with the inductance. Such a resistor will develop 1.16 watt-hours per megavolt-ampere per half cycle of current (60 cycles). This is the maximum energy which can be liberated per half cycle in the conventional linear R–L circuit on a steady-state basis in a 60-cycle circuit. The maximum inductive energy for this case is 0.3684 watt-hour per megavolt-ampere. Note that $1.16/0.3684 = \pi = 3.1416$. The maximum possible energy which can be liberated on a transient basis approaches the limiting value of 2.9473 watt-hours per megavolt-ampere per half cycle (exact value $= 1000/108\pi$). This is the maximum stored energy in the circuit inductance in a zero power circuit under full asymmetrical conditions. This is exactly eight times the maximum inductive energy in the conventional R–L circuit proportioned for maximum resistance wattages, that is, $2.9473 = 8 \times 0.3684$.

The arc energy liberated by the fuse is one of the primary ingredients contributing to the severity of its performance. The magnitude of the actual energy liberated is dependent upon the megavolt-amperes being interrupted. In general, the energy liberated increases as the megavolt-amperes increase; however, very important cases will be shown in which the energy actually decreases as the megavolt-amperes are increased.

Circuit Variables

Among the factors which influence the energy as well as the initial recovery voltage e_2 (the two primary ingredients of performance severity) may be listed the following:

1. Magnitude of the ratio of arc voltage to the circuit voltage, e_B/E_m.
2. Shape of the arc voltage.
3. Melting time of the fuse Θ_m.
4. The *MVA* available at the time of the short circuit.
5. Frequency of the circuit.
6. Power factor of the circuit.
7. Single-phase or 3-phase performance.
8. The displacement angle of the short circuit (Θ_1).

Only the first two of these factors are under the control of the fuse designer, and to some extent he is able to control the melting time Θ_m for any specific current rating. The last of these factors, however, is a purely random phenomenon. It behooves the fuse designer to have a thorough understanding of the influence of all eight of these factors, as well as their interrelation, in order to satisfactorily test fuses in the laboratory to meet all the requirements which may be encountered in their field application.

It is, of course, beyond the scope of this paper to present all the combinations and permutations in which these eight factors might occur to produce the most severe operating condition. Furthermore, no attempt will be made to wed the two most important elements of severity, arc energy k and recovery voltage e_2, into one severity factor inasmuch as these elements of severity seldom bear the same relative relationship for all fuses. Instead, it seems more practical at this time to show how both of these quantities vary by changing only one of the eight variables at a time. Some of the more exact quantitative values, as determined analytically, will be sacrificed in the interest of simplicity by an analytical treatment which approximates the actual fuse characteristics yet maintaining the same relative magnitudes and trends. A sufficient number of oscillograms of fuse performance are presented in part I of this paper to confirm the validity of the trends established by analytical means.

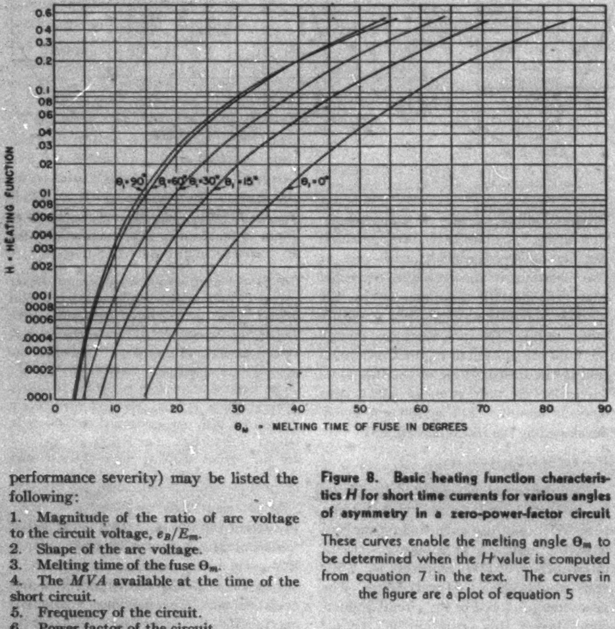


Figure 8. Basic heating function characteristics *H* for short time currents for various angles of asymmetry in a zero-power-factor circuit

These curves enable the melting angle Θ_m to be determined when the *H* value is computed from equation 7 in the text. The curves in the figure are a plot of equation 5

Effect of Arc Voltage—Generated Voltage Ratio

The single-phase performance of an interrupting device as effected by a variation in the magnitude of the applied voltage or in the magnitude of an assumed rectangular arc voltage e_B is presented in the five parts of Figure 3. Here rectangular arc voltages whose magnitudes represent 1.2, 1.0, 0.80, 0.50, and 0.10 times the crest of the generated voltage are presented. A discussion of Figure 3A already has been presented to enable a full understanding of the method used to determine the quantities presented on these diagrams. For the diagrams of Figure 3, the following factors are held constant at the values indicated:

1. Power factor of circuit assumed zero (Θ_p = 90 degrees).
2. Displacement angle of the short circuit (Θ_1) of 45 degrees which gives rise to an available per unit current peak I_m of 1.707.
3. Melting time of fuse (Θ_m) equal to 30 degrees thus causing fusion to take place at 75 degrees $=\Theta_s=(\Theta_1+\Theta_m)$.
4. Shape of arc voltage assumed rectangular.

On each diagram is given the arc energy coefficient k in watt-hours per megavolt-ampere as would be developed in a 60-cycle circuit. The quantity k_L represents the maximum stored energy in the circuit inductance. The magnitudes of peak current i_m as experienced by the interrupter during the interrupting process as well as the instantaneous recovery voltage e_2 at time of current zero are tabulated also.

The essential data from Figure 3 are plotted in Figure 4 as a function of the magnitude of the rectangular arc voltage. This shows that the full current-limiting effect of the fuse under these conditions is not realized unless the arc voltage exceeds the crest of the generated voltage. The initial recovery voltage e_2, an essential ingredient in determining the relative severity of the performance, is shown in such a way as to indicate not only the magnitude, but the direction of the generated voltage following interruption.

Although the data presented in Figure 4 represent the relative performance under the conditions as specified, it is representative enough to serve as a means of indicating roughly where the different types of circuit interrupters function with respect to the magnitude of their arc voltages. This is considered timely here in order that the reader may see in what position of the over-all interrupting spectrum a fuse of the current-limiting type lies, and at the same time to point out that the analytical methods presented are applicable to any type of circuit interrupter. The bar indications at the top of Figure 4 thus show approximately the

Table I

$\frac{I}{1,000}$	H	Θ_m	k	MVA	W
3	2.78	103.3	2.36	7.5	17.7
4	1.57	86.2	2.18	10.0	21.8
5	1.00	75.7	1.93	12.5	24.1
10	0.2507	50.5	1.22	25.0	30.5
15	0.1114	40.3	0.84	37.5	31.5
17.5	0.082	37.3	0.73	43.75	31.92
20	0.0627	33.9	0.63	50.0	31.5
25	0.0401	29.8	0.44	62.5	27.5
30	0.0278	26.6	0.347	75.0	26.0
40	0.0157	22.4	0.205	100.0	20.5
50	0.0100	19.6	0.132	125.0	16.5
60	0.00697	17.6	0.093	150.0	14.0
80	0.00392	14.8	0.053	200.0	10.6
100	0.002507	13.0	0.036	250.0	9.0

relative arc voltages associated with the following types of circuit interrupters.

1. Air blast circuit breakers.
2. Oil circuit breakers.
3. Expulsion type fuses.
4. Magnetic type air circuit breakers.
5. Current-limiting fuses.

The curves of Figure 4, when considered in this light, hence represent only relative values and describe the arcing period only. In a circuit breaker, Θ_m represents the contact parting time as measured from the previous current zero.

Because arc voltage is expressed, in this paper, in per unit of generated voltage crest, it follows that the diagrams of Figures 3 and 4 also represent the change that takes place as the generated voltage is increased, which in turn is equivalent to decreasing the arc voltage for a given fuse. Such use of these diagrams demonstrates the flexibility of the use of the per unit system adopted. It also follows that if both *k* and the circuit megavolt-amperes are increased as the generated voltage is increased, then the actual arc energy (*k*×megavolt-amperes) will increase faster than the increase in *k* alone.

Effect of Displacement Angle, Θ_1

The arc energy coefficient curve *k* on Figure 4 is replotted in Figure 5 together with similar curves representing three other displacement angles, keeping the melting angle fixed at 30 degrees. These curves describe, better than words, the relative arc energies as the displacement angle is varied. For the low arc voltage interrupters, the relative arc energies are progressively lower as the fault starting angle is increased. For high arc voltages (above 1.2) the liberated energy is reversed from that of the low arc voltage devices. In the upper right-hand corner of Figure 5 are shown the relative arc energies plotted with respect to Θ_e or $(\Theta_1+\Theta_m)$ for a specific arc voltage of 1.2. This shows that the peak energy released occurs when the current-responsive element melts just prior to the peak of the generated voltage. This fact has been determined empirically by other investigators,[7] but is subject to qualifications which will develop under other headings.

In the testing of fuses, therefore, the choice of a displacement angle which permits the current-responsive element to melt prior to the generated voltage crest will be, in general, one of the regions of severity which should be explored to demonstrate satisfactorily the adequacy of the fuse to meet its rating.

Arc Voltage Wave Shape

Seldom does the arc voltage of a fuse follow the pattern of a rectangular wave shape, although this simple assumption serves adequately to determine analytically the relative trends in fuse severity. The arc voltage e_B, shown in Figure 6 is more representative of one type of arc voltage found in the current-limiting fuses. Because the power factor still is assumed to be zero ($\Theta_p=90$ degrees), it follows that the areas *A* and *B* still must be equal from current zero to current zero. The cancellation current i_B resulting from the arc voltage is no longer a straight line, but is dependent upon the volt–time area under the arc voltage. The quantities in the diagram of Figure 6 were so chosen as to place the current interruption Θ_2 at the same point as occurred in Figure 3B for the rectangular arc voltage whose magnitude was $1.0E_m$. Hence the initial recovery voltage in both Figures 6 and 3B is $-0.10E_m$. The energy liberated by the fuse in the case of Figure 6, however, is only 70 per cent of that for the case shown in Figure 3B. This apparently makes the drooping arc voltage of Figure 6 a more desirable characteristic. This is true of the two cases shown. However, conclusions on this point should be reserved until the effect of circuit frequency has been covered in the next paragraph.

Circuit Frequency Versus Arc Voltage Wave Shape

Table II

I	H	Θ_m	k	MVA	W
500	0.437	50.5	1.49	1.25	1.86
800	0.1713	45.0	1.03	2.00	2.06
1,000	0.1092	40.2	0.84	2.50	2.10
1,100	0.0906	38.3	0.77	2.75	2.11
1,200	0.0760	36.4	0.70	3.00	2.10
1,400	0.0560	33.2	0.58	3.50	2.03
1,700	0.0379	29.1	0.433	4.25	1.84
2,000	0.0274	26.3	0.335	5.00	1.68
2,500	0.0176	23.1	0.227	6.25	1.42
3,000	0.0122	20.7	0.158	7.50	1.19
4,000	0.00684	17.5	0.091	10.00	0.91
5,000	0.00438	15.3	0.059	12.50	0.738

When the fuse which has the arc voltage characteristic shown in Figure 6 is applied to a zero power factor 25-cycle circuit (Figure 7) and a fault is initiated at $\Theta_1=45$ degrees as in Figure 6, the current-responsive element will melt in 22.5 degrees instead of the 30 degrees as in the 60-cycle circuit. This is to be expected and is shown thus in Figure 7 to give a fair comparison between the two frequencies under the conditions assumed. The arc voltage of Figure 6, however, when applied to the 25-cycle scale of Figure 7, drops so rapidly that the area *B* is not large enough to neutralize the area *A*, and accordingly the arc current *i* stops its descent and again starts to rise and continues to do so as long as the arc voltage e_B is below the generated voltage. Peak arc current then occurs when the area A' is more than *B* and takes place as the arc voltage again crosses the generated voltage. The magnitude of the peak arc current is then equal to volt–time areas $A+A'-B$ divided by the circuit inductance. (Inductance is unity in the per unit system.) Current zero eventually occurs when the area B' has been supplied such that $A+A'=B+B'$. The energy developed by this performance in a 25-cycle circuit is $k=1.56$ watt-hours per megavolt-ampere or 2.62 times the energy represented in Figure 6 for the same arc voltage acting in a 60-cycle system of the same *megavolt-ampere*. The severity of the performance shown in Figure 7, however, is further complicated by the high positive value of the initial recovery voltage e_2 of $+0.78\ E_{max}$. Such a combination of high energy and high recovery voltage would make interruption difficult. A study of Figures 6A and 8B of part I of this paper[1] will confirm the validity of the analysis shown in Figure 7 of this paper.

Had the arc voltage of Figure 3B been applied to the 25-cycle conditions of Figure 7, the arc energy coefficient *k* still would have been in the neighborhood of 1.5, but the initial recovery voltage e_2 would have been leading at about $-0.25\ E_m$. One of the factors favoring the 25-cycle system which tends to compensate for the apparently high arc energy is the fact that the recovery voltage rises much slower and the peak is reached much later. This advantage can be enjoyed, however, only when interruption occurs in the neighborhood of voltage zero. This analysis explains in part why heavy cur-

Table III

Starting angles, Θ_1	0°	30°	60°	90°
Arc voltage angle Θ_e for max energy, as shown in Figure 10	67.5°	68.0°	78.5°	112.5°

rent fuses usually are derated when applied to 25-cycle circuits.

Determination of Melting Time

The performance of all high-speed interrupting devices are affected by the magnitude of the circuit current at the time the arc voltage is introduced into the circuit. This is particularly true of those interrupters which must permit only one pulse of arc current before interruption. In a circuit breaker, this interval is determined by the contact parting time and this time can be predetermined accurately only in very high-speed devices.

In a fuse, however, this interval is determined by the fusion time of the fusible element and accordingly may be predetermined quite accurately for large currents from the design data of the fuse. When the currents are large and the fusion times are less than one cycle, it can be shown that the total i^2t absorbed by the fuse is the true criterion of its fusion characteristics regardless of the manner in which this i^2t is introduced into the circuit. This fact is predicated upon the assumption that in this short interval, no heat is dissipated from the fusible element. It follows, therefore, that the melting time of the fuse may be predetermined if the i^2t of the short-circuit current is available for all conditions which may be experienced. This has been determined in Appendix II in such a way as to be applicable for all currents and all frequencies. Here it is shown that

$$\sum i^2t = \frac{H \times I^2}{\omega} \tag{4}$$

where

Σi^2t is known from the fuse data (equation 7)

I = the available symmetrical value of the zero power factor short-circuit current being considered

$\omega = 2\pi f$

f = circuit frequency

H = a dimensionless trigonometric heating function whose general equation for the single phase zero power factor circuit is expressed as

$$H = 2\Theta_m + \Theta_m \cos 2\Theta_1 + {}^1/_2 \sin 2\Theta_e + {}^3/_2 \sin 2\Theta_1 - 4 \cos \Theta_1 \sin \Theta_e \tag{5}$$

This heating function is extremely useful and is plotted for various fault starting angles Θ_1 using as the abscissa the melting time Θ_m of the fuse. The melting angle of any fuse can be determined therefore, by first determining H from

$$H = \frac{\Sigma i^2t \times \omega}{I^2} \tag{6}$$

and in turn referring to the H function curves to determine the melting angle Θ_m which in turn may be converted to time if desired.

For a multiconductor fuse, the value of Σi^2t is given by

$$\sum i^2t = F_m \times d^4 \times n^2 \tag{7}$$

where

n = number of conductors in multiple

d = their diameter in mils

F_m = a constant depending upon the metal of the fuse conductor

Effect of Melting Time of the Fuse

From the heating function characteristics H shown in Figure 8, it is observed that as H is decreased by raising the short-circuit current I, the melting time Θ_m decreases. It is observed also from Figure 9 that as the melting time Θ_m is decreased, the magnitude of arc energy function k is decreased; although the megavolt-amperes (by which k is multiplied to produce the actual arc energy) increases proportional to the current. Because the melting time decreases as the current is raised, the question naturally arises concerning the relationship between the actual liberated energy W and the symmetrical fault current, as the latter is increased without limit for a particular fuse design. This problem is approached in the following text for a fuse which has an Σi^2t of 66,500.

By selecting a single-phase generated voltage of 2,500 volts and assuming a per unit rectangular arc voltage of 1.2 or 4,250 volts and further selecting the closing angle Θ_1 of the zero power short circuit at 30 degrees for each current applied, these curves are entered to obtain the solution which is tabulated in Table I for a 60-cycle circuit.

From Table I it becomes apparent that for the assumed condition of a flat top or rectangular arc voltage of 1.2 times the crest of the generated voltage together with a closing angle of 30 degrees in a 60-cycle zero power factor circuit, the maximum energy absorbed by the fuse (under these conditions) will occur in the neighborhood of 17,500 amperes. The energy developed under these conditions will be 31.92 watt-hours. As the fault current is increased above 17,500 amperes, the developed energy decreases until at 100,000 amperes only 9.0 watt-hours of energy is available, the arc voltage, of course, being assumed constant at 1.2 times the crest of the generated voltage as the current is increased.

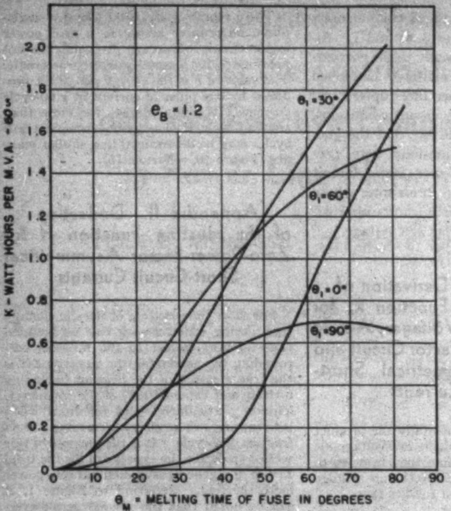
This problem is presented again in Table II for a smaller fuse having an Σi^2t of 290 and for the same assumed conditions as for the preceding case.

The two arc wattage characteristics W tabulate in Tables I and II are plotted in the master curve of Figure 10 as the $\Theta_1 =$ 30-degree curve together with the corresponding current–energy variations for fault starting angles of $\Theta_1 = 0$ degrees, $\Theta_1 = 60$ degrees, and $\Theta_1 = 90$ degrees. The curves of Figure 10 apply for any fuse operating in a zero power factor circuit at any frequency, but assumes a rectangular arc voltage of 1.2 times the crest of the generated voltage. The fact that one such master energy curve exists for all fuses for all frequencies for a specific arc voltage ratio e_B, is a natural consequence of the choice of parameters which made possible this interesting observation.

Figure 10 teaches that for an assumed arc voltage of 1.2 times the crest of the generated voltage, maximum energy conditions occur for a fault starting angle of $\Theta_1 = 0$ degrees. Similar master characteristics for other arc voltage-generator voltage ratios e_B, shown in the discussion of this paper, reveal that the fully displaced fault does not always give rise to the maximum possible energy. For example, when the ratio e_B is 1.0 instead of 1.2, maximum energy conditions are indicated for a fault starting angle of about 50 degrees and the energy levels are approximately 50 per cent higher than for the e_B ratio of 1.2.

As the rectangular arc voltage e_B drops below 1.0 to, say, 0.95 or 0.90 E_m and below, the arc energy characteristic as a function of megavolt-amperes no longer decreases for high available fault currents, but rises with the megavolt-amperes. The tendency for the arc voltage of the fuse to decrease with increasing circuit current hence may bring about a rising arc energy characteristic at very high currents, and thus establish a ceiling on the interrupting capacity of the fuse. The critical angle of such a characteristic occurs for a Θ_1 between 50 and 70 degrees. A slow rise of the arc voltage of the fuse will contribute materially to this phenomenon, particularly for fuses of the low continuous ampere capacity. These conclusions may be confirmed easily by the use of the equations or graphical methods presented in this paper.

Of considerable interest is the fact that,



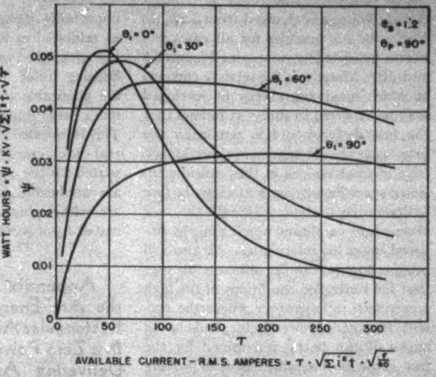


Figure 10 (above). Master energy characteristics for fuses, as a function of single-phase symmetrical zero-power-factor short-circuit currents

These relations are valid only for rectangular arc voltages (or their equivalent) with a nominal magnitude of 1.2 times the crest of the generated voltage. They may be used for any frequency or applied voltage under these conditions. See equation 13

Figure 9 (left). Typical arc energy curves for a 60-cycle zero-power-factor system computed for a rectangular arc voltage which is 1.2 times the crest of the generated voltage

as shown in Figure 10, maximum available energy does not occur necessarily at maximum available rms current. The peak energies of Figure 10 occur when the arc voltage angles are as shown in Table III for $e_B = 1.2$ and $\Theta_p = 90$ degrees.

This confirms the empirical observation made in part I of this paper and reference 7 that one region of maximum severity occurs when melting of the fusible element occurs between 70 degrees and 80 degrees of the voltage wave, that is, just prior to the voltage peak (see Figure 5). This feature is discussed more fully in part I of this paper.

It also is revealed in Figure 10 that the relative energies to be absorbed by fuses of various sizes increase as the square root of the $\Sigma i^2 t$ and directly with the voltage of the application, provided that the arc voltage ratio e_B is maintained constant.

Effect of Circuit Frequency

Figure 10 also teaches that the maximum energy conditions for various circuit frequencies vary as the square root of the ratio of the frequencies. For example, in a 25-cycle circuit, the increase energy available to be absorbed by the fuse, under the conditions of maximum severity, would be $\sqrt{60/25}$ or 1.55 times that for a 60-cycle circuit. It is imperative that the assumption under which this is true be kept firmly in mind; namely, that in both cases, a rectangular arc voltage of the same relative magnitude is maintained throughout the arcing time of the fuse. In practice, the possible energy ratios for the two frequencies exceeds 1.55 because of the phenomena discussed in connection with Figure 7.

Circuit Power Factor

It is beyond the scope of this paper to include equations and curves which expand this treatment to circuits of various power factors. Such an extension has been indicated clearly,[15,17] and although slightly more complicated, should result in curves similar to those presented in this paper. A few comments regarding the broad conclusions drawn from calculations made for circuits of various power factors are:

1. In general, if the short-circuit calculations are made neglecting all resistance, regardless of its presence, and circuit interrupters are applied which are capable of interrupting the zero power factor megavolt-ampere so determined, the arc energy released, in general, will be less than the values indicated in this paper. On this basis the applications will be on the conservative side.

2. If the short-circuit determinations are made by test, the actual power factors also should be available and the above zero power factor megavolt-ampere readily determined.

3. If the applications of interrupters are made up to their full rated megavolt-amperes, using currents determined from test or calculation which include sufficient resistance to produce currents which lag the voltage by 60 to 75 degrees, then the arc energy released by high speed interrupters, especially of the current-limiting type, will be increased above those indicated by the curves and equations in this paper.

(a). Devices which operate on the right hand side of the master energy characteristics of Figure 5, that is, high arc voltage devices of a high speed nature, will be more likely to experience an increase in circuit severity under these conditions.

(b). High speed devices which operate on the left-hand side of the master energy characteristics of Figure 5, that is, low arc voltage devices, will experience a small increase in developed energy over the zero power factor circuit (constant megavolt-amperes). Such interrupters, however, usually are sensitive to the recovery voltage transient and must consider this voltage in their severity problem. For this reason such interrupters are likely to experience an over-all reduced severity resulting from the damping of the recovery voltage transient offered by the presence of the circuit resistance.[18] This feature will be modulated to a considerable extent depending upon the natural frequencies of the recovery voltage oscillation associated with the circuit.[16] The performance of expulsion fuses is likely to fall into this classification.

Discussion

It has been possible in the preceding treatment to discuss the effect of only one circuit variable at a time. It will become clear to the reader that when, for example, the arc voltage is changed from $1.2E_g$ to $1.0E_g$, the arc energies for all cases will rise as indicated by comparing Figures 3A and 3B. Moreover, the relative currents at which maximum energy is developed may be expected to shift. It follows that the fuse designer and in particular the fuse tester should have a working knowledge of what regions in this multidimensional space he may expect to test the fuse to determine its ability to handle the most severe field condition which may be imposed upon his interrupter. It also will become apparent to the ultimate fuse user that the controlled conditions of the high power testing laboratory where the current, frequency, power factor, and closing angle all can be predetermined for the test, is the most reliable means of proving the adequacy of power fuses to meet all field conditions.

It should be re-emphasized here that the curves and relationships shown are predicated upon assumed magnitudes of arc voltage of a fuse and do not represent a tabulation or plot of any particular fuse data. In fact, no fuse to our knowledge will produce exactly a rectangular arc voltage. For these reasons, the results obtained on this analytical basis can be used only as a guide to the trends which more or less do take place as described in part I of this paper. Experience shows that as the current at melting increases, the magnitude of the arc voltage tends to decrease. Hence we have what is akin to a volt–ampere characteristic for a particular fuse. With such a characteristic established, the true severity ratios for a particular fuse might be determined more accurately by the application of the curves presented in this paper.

The reader is directed particularly to part I of this paper where oscillograms will be found which demonstrate the principles presented in this paper. The nomenclature on the oscillograms is identical to that used in both papers. The volt–time area relationships are indicated clearly.

An analytical approach to the problem of current-limiting fuse performance has been presented which suggests a method of attack which holds much promise of extension to conditions not presented in this paper. The method has a firm foundation on fundamental principles of circuit theory. Only a little region of a multidimensional circuit problem has been covered in this paper, but sufficient to indicate the essential trends in a quantitative manner for a particular interrupter. The use of superposition principles affords the opportunity of a clear geometrical visualization of the relative importance of each of the circuit variables as exhibited in Figures 3, 6, and 7. Although applied specifically to the current-limiting fuse, the methods employed are applicable to other interrupters and for a wide range of circuit conditions. This is revealed by the fact that the general arc energy equations here presented for the zero power factor circuit are applicable to any interrupter whose arc voltage characteristic can be approximated by a rectangular arc voltage.

Appendix I. Derivation of the Arc Energy Function *k*, for Rectangular Arc Voltages, Acting in a Zero Power Factor Circuit and Delivering Asymmetrical Short-Circuit Currents

Arc energy is fundamentally the product of the arc current i and the arc voltage e_B, integrated over the arcing time from Θ_s to Θ_2 (see Figure 3). Because the arc voltage is to be held constant, it follows that

$$\text{arc energy} = e_B \int_{\Theta_s}^{\Theta_2} i\, d\Theta \qquad (8)$$

This integral will be observed to represent the area between I_{sc} and i_B between Θ_s and Θ_2 in all cases, as shown in Figure 3. Inspection will reveal this area to be represented by

$$\int_{\Theta_s}^{\Theta_2} (\cos\Theta_1 - \cos\Theta)\,d\Theta - {}^1/_2\Theta_B(\cos\Theta_1 - \cos\Theta_2) \qquad (9)$$

The solution becomes

$$\int_{\Theta_s}^{\Theta_2} i\,d\Theta = [{}^1/_2\Theta_B(\cos\Theta_1 + \cos\Theta_2) - (\sin\Theta_2 - \sin\Theta_s)] \qquad (10)$$

To obtain the true magnitudes of current the foregoing instantaneous per unit value must be multiplied by $\sqrt{2}I$. Similarly the per unit arc voltage e_B must be multiplied by $\sqrt{2}E_g$ to give the actual arc voltage. Moreover, because $\Theta = \omega t$, it is necessary to divide the preceding integral by ω to convert the area to ampere-seconds. Making these additions and at the same time allowing the foregoing bracketed term to be represented by the symbol Q, there results

$$\text{arc energy} = \frac{2 \times E_g \times I \times e_B \times Q}{\omega} \text{ watt-seconds} \qquad (11)$$

Converting to watt-hours per megavolt-ampere, this becomes

$$\text{arc energy} = 555\,\frac{e_B \times Q}{\omega} \text{ watt-hours per megavolt-ampere} \qquad (12)$$

For the 60-cycle circuit, where $\omega = 377$, this energy equation becomes

$$k = 1.4736 e_B \times Q \text{ watt-hours per megavolt-ampere (60 cycles)} \qquad (13)$$

where

$$Q = [{}^1/_2\Theta_B(\cos\Theta_1 + \cos\Theta_2) - (\sin\Theta_2 - \sin\Theta_s)]$$

These relations are valid for any single-phase interrupter acting in a zero power factor circuit whose arc drop may be approximated by a rectangular characteristic. The value of k is the energy per *MVA* produced by one pulse of current in a 60-cycle current. If arcing persists for more than one-half cycle, the energy in subsequent half cycles may be determined in a similar manner (Figure 10, reference 15).

Appendix II. Derivation of the Heating Function *H* for Zero-Power-Factor Asymmetrical Short-Circuit Currents

For short time heating of electric conductors, during which time it can be assumed that no heat is lost to the medium surrounding the conductor, the quantity Σi^2t is the true measure of the heating (including melting and vaporization) of the conductor. This is particularly valid for fuses when interrupting short-circuit currents in a fraction of a cycle. It is the purpose of this section to derive the expression for the total Σi^2t of the general asymmetrical zero power factor current. Referring to Figure 1, it can be noted that the general zero-power-factor short-circuit current may be expressed

$$I_{sc} = \sqrt{2}I(\cos\Theta_1 - \cos\Theta) \qquad (14)$$

It follows that the total or summation Σ of i^2t over the melting interval may be obtained from the integration of $i^2\Theta$ between the limits Θ_1 and Θ_s; namely

$$\sum i^2\Theta = 2I^2 \int_{\Theta_1}^{\Theta_s} (\cos\Theta_1 - \cos\Theta)^2\, d\Theta \qquad (15)$$

The solution of this equation is

$$\sum i^2\Theta = I^2\{2\Theta_m + \Theta_m \cos 2\Theta_1 + {}^1/_2 \sin 2\Theta_s + 1{}^1/_2 \sin 2\Theta_1 - 4\cos\Theta_1 \times \sin\Theta_s\} \qquad (16)$$

where

$$\Theta_m = (\Theta_s - \Theta_1)$$

Allowing the bracketed trigonometric function of equation 16 to receive the symbol H, there results

$$\sum i^2\Theta = I^2H \qquad (17)$$

But $\Theta = \omega t$ (radian measure), hence

$$\sum i^2t = \frac{I^2H}{\omega} \qquad (18)$$

and

$$H = \frac{\Sigma i^2t \times \omega}{I^2} \qquad (19)$$

H is termed the zero power factor heating function and is computed from equation 16 and plotted in Figure 8 as a function of the melting angle of the fuse in degrees for various fault starting angles, Θ_1. Because the Σi^2t is available from the fuse design (see equation 7), and the circuit frequency and symmetrical short-circuit current are known, it follows that the melting angle of any fuse can be determined from Figure 8 or its extension by the use of equations 16 and 22. The presence of ω (which is $2\pi f$) in these rela-

The Use of Business Machines in Determining the Distribution of Load and Reactive Components in Power Line Networks

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THE ANALYZER BOARDS available at various points in the United States have afforded welcome relief from the tedious calculations which are ordinarily necessary in determining the distribution of load and reactive component in electric power networks. Since, however, they frequently are located inconveniently, power engineers still have to use longhand methods on their current problems and often have to content themselves with approximate methods, yielding less information than they would like to have. Help in the solution of these problems that arise between visits to the analyzer board can be had by making use of accounting machines of the IBM* type, which usually are available to members of the light and power industry. These machines greatly expedite the calculations while retaining full accuracy of the classical method.

AIEE defines network[1] as follows:

> "An electric network is a system of interconnected admittances.
>
> The point at which three or more admittances meet is called a junction point of the network. An admittance connecting any two junction points is called an arm or branch of the network. A mesh within a network is a closed path through three or more junction points."

For the purpose of this paper we will use the expression mesh in the sense of the foregoing definition but with the limitation that a mesh will be in its simplest form, that is, containing no smaller meshes.

The determination of the distribution of load and reactive component in the branches of a network can be accomplished by the solution of a group of simultaneous complex linear equations, with one equation for each mesh of the network; but the large amount of labor involved ordinarily makes this procedure impracticable for networks involving more than three or, at the most, four meshes. Simultaneous equations are solved most easily by the method of determinants. IBM machines are used in the solution of the determinants since they are designed specially to automatically carry through schematic sequences of operations involving multiplication, addition, and subtraction.

For reasons of simplification, a typical 2-mesh network, requiring a determinant of the second order, is shown in Figure 1. Power is fed into the network by two generating stations, 1 and 2, and power is taken out from the network at points *a*, *b*, and *c*. It is assumed that generating station 2 is operating at a fixed load and substantially fixed excitation. Generating station 1 is doing the load regulating.

The vector load currents $\boldsymbol{I}_a$, $\boldsymbol{I}_b$, and $\boldsymbol{I}_c$ and generator current $\boldsymbol{I}_d$ are known with respect to a common reference axis, which in this case is assumed to be the potential $\boldsymbol{E}_0$ (which also is known) at generating station 1. The given vector impedances of each of the branches of the network are represented by the $\boldsymbol{Z}$ values shown in the diagram.

The key to the problem is to determine the two unknown branch currents $\boldsymbol{I}_1$ and $\boldsymbol{I}_2$ in branches *o–f* and *f–d*, respectively, from which the complete solution readily follows. It will be observed from the equations of the currents of the other five branches that they are all functions of known currents and one or both of the two unknown currents $\boldsymbol{I}_1$ and $\boldsymbol{I}_2$.

Referring to Figure 1, the usual solution of a typical 2-mesh network is carried through by means of a second order determinant, as shown in equations 1 to 13. The end product of this procedure is represented by equations 11, 12, and 13, which involve the real and imaginary *M* and *N* quantities used directly in the solution of the values $\boldsymbol{I}_1$ and $\boldsymbol{I}_2$. The right hand members of equations 8, 9, and 10 are developed by the determinant routine, and it is with these that the machines deal.

For example, equation 8 consists of eight terms of two factors each, four of these are real and four of them are imaginary terms. The machines obtain the products of the factors in each term and the final algebraic summation of the terms, keeping the real terms and the imaginary terms separate.

In the case of a 3-mesh network, in which three simultaneous equations are required, each determinant has 48 terms of three factors each. In general the order of the determinant will be the same as the number of meshes in the network

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* International Business Machines.

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tions permits their use for any frequency. The determination of Θ_m is necessary to determine the arc energy of the interrupter from Figure 9 and derived in Appendix I.

Equation 19 is the foundation of the abscissa of Figure 10 in the observation that

$$I=\sqrt{\Sigma i^2 t}\sqrt{\omega}\times\frac{1}{\sqrt{H}} \tag{20}$$

$$I=\sqrt{\Sigma i^2 t}\sqrt{\frac{f}{60}}\times\sqrt{\frac{377}{H}} \tag{21}$$

where in Figure 10,

$$\tau=\sqrt{\frac{377}{H}}$$

Approximation Formula

When the melting angles are very small (up to 0.1 radian) the bracketed portion of equation 16 reduces to the approximate formula

$$H\cong\left\{{}^2/_3\sin^2\Theta_1\times\Theta_m{}^3+{}^1/_4\sin 2\Theta_1\times\Theta_m{}^4+{}^1/_{10}\cos^2\Theta_1\times\Theta_m{}^5\right\} \tag{22}$$

For the symmetrical case, when $\Theta_1=\pi/2$, this reduces to

$$H\cong{}^2/_3\Theta_m{}^3 \tag{23}$$

For the asymmetrical case, when $\Theta_1=0$, the heating function becomes

$$H\cong{}^1/_{10}\Theta_m{}^5 \tag{24}$$

These relations are quite useful when the currents are high enough to cause melting in a few degrees.

References

(See page 1034.)

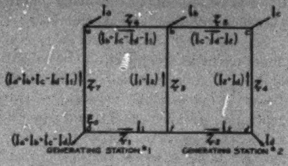

Figure 1. Typical 2-mesh network

and any complex determinant of the *n*th order will have ($2^n\, n!$) terms of *n* factors each.

Experience shows that the solution of a 4-mesh network by the usual method takes from three to ten days, depending on whether or not any errors are made in the 7,680 calculations and 9,600 determinations of signs. The machines will do the same work in four to six hours, requiring only the attention of the regular machine operator. The only errors that can arise in the machine method will be caused by supplying incorrect coefficient data to the master cards(subsequently described). This can be avoided by being careful that the coefficients are correct. This part of the work is not difficult or lengthy, but requires careful checking. If no preliminary work has been done, it will take approximately one day to supply the coefficients for a 4-mesh network. This part of the calculation is the same for the usual method and the machine method.

The process of solving complex determinants by means of the machines is essentially as follows: The accounting machine people punch a so-called formula deck for solving a particular order of determinant. Each factor will be represented by a card in this deck for each term in which the factor appears. For example, in a determinant of order 2 there are eight factors, each of which appears in two terms, making 16 cards in all. A determinant of the fourth order has 32 factors, each of which appears in 48 terms, requiring 1,536 cards; and a determinant of the fifth order has 50 factors, each appearing in 384 terms, requiring 19,200 cards. In addition there will be almost as many blank cards as formula cards. These are punched automatically in the calculating process, one for each term of the determinants. The cost of the cards is nominal. An examination of equation 8 will show substantially what is accomplished by the formula deck. Each card of the deck is punched with certain coded signal numbers that indicate to the various machines the process to be followed. In equation 8 the formula cards will start the sequence of operations required by the determinant, as $a_1 \times a_4$, $b_1 \times b_4$, $a_2 \times a_3$, and so forth, and will signal the sorting machine in its turn to sort the real values separately from the *j* values. In this way the tabulating machine can make the proper additions and subtractions of terms to give the *M* and *N* values of equations 11, 12, and 13.

The formula deck is, in reality, a signal deck capable of producing a definite sequence of multiplications and additions and is not concerned with numerical values. It will solve any determinant of the order for which it was designed, provided the numerical values are assigned. These values for any particular problem are assigned to the formula deck by the use of so-called master cards, on which have been punched the numerical values of the real and *j* coefficients of the fundamental equations 3 and 4. A determinant of the fourth order requires 32 of these cards, and a determinant of the fifth order requires 50 cards. The master cards are arranged in the formula deck in a definite order, and all the cards then are run through a machine which punches on the cards of the formula deck the numerical values carried by the master cards. Duplicates of the original formula deck are produced automatically by machine, and these duplicates are used in the solution of particular problems, the original deck being used over and over in this way.

The duplicate formula deck, with numerical values assigned, then is run through the various machines for sorting, multiplication of factors, summation of terms, and so forth. Our information indicates that the machines will handle determinants up to the fifth order at least, and their use therefore makes practicable the solution of many network problems which otherwise would have to wait until they could be set up on an analyzer board.

The detailed procedures involved in hand punching, machine punching, and running the cards have not been discussed since all of these are matters of routine with machine operators. The purpose of this paper merely is to call attention to the adaptability of the machines for the solution of complex determinants using a method developed for the Puget Sound Power and Light Company by the Seattle office of IBM Company.

Appendix

Given

I_a, I_b, and I_c = vector load currents
I_d = vector generator currents
E_g = vector generator potential
Z_1, Z_2, and so forth = vector impedance operators

$$I_0 = I_a + I_b + I_c - I_d$$

The problem is to find the vector line currents I_1 and I_2.

From Kirchhoff's laws

$$I_1Z_1 + (I_1 - I_2)Z_6 = (I_a + I_b + I_c - I_d - I_1)Z_7 + (I_b + I_c - I_d - I_1)Z_8 \quad (1)$$

$$(I_1 - I_2)Z_6 + (I_c - I_d - I_2)Z_5 = I_2Z_2 + (I_2 + I_d)Z_4 \quad (2)$$

Collecting coefficients of $_1$ and I_2 in equations 1 and 2 we have

$$I_1(a_1 + jb_1) + I_2(a_2 + jb_2) = c_1 + jd_1 \quad (3)$$

$$I_1(a_3 + jb_3) + I_2(a_4 + jb_4) = c_2 + jd_2 \quad (4)$$

where

$$a_1 + jb_1 = A_1 = Z_1 + Z_6 + Z_8 + Z_7$$
$$a_2 + jb_2 = B_1 = -Z_6$$
$$c_1 + jd_1 = Q_1 = I_aZ_7 + (I_b + I_c - I_d)(Z_8 + Z_7)$$
$$a_3 + jb_3 = A_2 = Z_6$$
$$a_4 + jb_4 = B_2 = -Z_2 - Z_6 - Z_4 - Z_5$$
$$c_2 + jd_2 = Q_2 = I_d(Z_4 + Z_5) - I_cZ_5$$

Putting the equivalent coefficients of I_1 and I_2 from equations 3 and 4 into determinant form,

$$S_0 = \begin{matrix} A_1 & B_1 \\ A_2 & B_2 \end{matrix} = A_1B_2 - A_2B_1 \quad (5)$$

$$S_1 = \begin{matrix} Q_1 & B_1 \\ Q_2 & B_2 \end{matrix} = Q_1B_2 - Q_2B_1 \quad (6)$$

$$S_2 = \begin{matrix} A_1 & Q_1 \\ A_2 & Q_2 \end{matrix} = Q_2A_1 - Q_1A_2 \quad (7)$$

$$I_1 = S_1/S_0$$
$$I_2 = S_2/S_0$$

Then equations 5, 6, and 7 could be written, respectively, as follows:

$$S_0 = (a_1 + jb_1)(a_4 + jb_4) - (a_3 + jb_3)(a_2 + jb_2) = (a_1a_4 - b_1b_4 - a_2a_3 + b_2b_3) + j(a_1b_4 + a_4b_1 - a_3b_2 - a_2b_3) \quad (8)$$

$$S_1 = (c_1 + jd_1)(a_4 + jb_4) - (c_2 + jd_2)(a_2 + jb_2) = (a_4c_1 - b_4d_1 - a_2c_2 + b_2d_2) + j(a_4d_1 + b_4c_1 - b_2c_2 - a_2d_2) \quad (9)$$

$$S_2 = (c_2 + jd_2)(a_1 + jb_1) - (c_1 + jd_1)(a_3 + jb_3) = (a_1c_2 - b_1d_2 - a_3c_1 + b_3d_1) + j(a_1d_2 + b_1c_2 - b_3c_1 - a_3d_1) \quad (10)$$

From equations 8, 9, and 10 let

$$a_1a_4 - b_1b_4 - a_2a_3 + b_2b_3 = M_0$$
$$a_1b_4 + a_4b_1 - a_3b_2 - a_2b_3 = N_0$$
$$a_4c_1 - b_4d_1 - a_2c_2 + b_2d_2 = M_1$$
$$a_4d_1 + b_4c_1 - b_2c_2 - a_2d_2 = N_1$$
$$a_1c_2 - b_1d_2 - a_3c_1 + b_3d_1 = M_2$$
$$a_1d_2 + b_1c_2 - b_3c_1 - a_3d_1 = N_2$$

$$S_0 = M_0 + jN_0 \quad (11)$$

$$S_1 = M_1 + jN_1 \quad (12)$$

$$S_2 = M_2 + jN_2 \quad (13)$$

Reference

1. DEFINITIONS OF ELECTRICAL TERMS. American Standard C42, 1941. Section 05.45.080.

Dielectric Strength of Station and Line Insulation to Switching Surges

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THE STRIDES achieved since the early thirties in the field of system-insulation rationalization are well known in recent technical literature[1-7] and need not be reviewed here. However, a point to emphasize is that further progress in this industry-wide problem can be attained only through continued research and studies directed particularly to the remaining major questions. In this manner it will be possible to pursue in future developments the principles of system-insulation rationalization to their fullest economic possibilities.

A survey of these future developments would indicate the need for a greater knowledge of the conditions resulting from system transients and, correspondingly, more comprehensive data on the performance of insulation to these conditions. Much has been published on system overvoltages, and there have been some contributions[8-11] relating to the strength of insulation to switching surges and other transients, but the fact remains that more comprehensive data are required in view of the future developments. It is the purpose of this paper to fill in the existing gap by presenting recently established data on the strength of station and line insulation to switching surges, and also to discuss and analyze the new findings with previously determined data in relation to the over-all problem.

Importance of Switching-Surge Data

It would be well to review briefly the significance of system overvoltages in relation to the insulation requirements. This may be done considering several representative cases in which switching surges appear to assume particular importance. Insulation may be proportioned on circuits not exposed to lightning with reference to the system overvoltages entirely, of course, in addition to the normal operating voltages. The data presented here therefore should be helpful in applications of this character.

For systems exposed to lightning, it is a fact that the insulation is applied primarily from a lightning or impulse viewpoint and, in so doing, the requirements arising from system overvoltages as well as other requirements usually are fulfilled, or more than fulfilled, simultaneously. Almost invariably this is the case for systems of modern design which operate with the neutral solidly grounded. However, this approach cannot be accepted

C_s
G
B
C_l
CD
CRO
L
R
G
C_l
C_s

Figure 1. General view of and circuit arrangement for switching-surge tests on medium voltage apparatus

C_s—Surge-voltage generator, 0.083 microfarad
C_l—Load capacitance, 0.042 microfarad
R—Load resistance, 384,000 ohms
L—Series inductance, 1.06 henries
G—25-centimeter sphere gap
B—Apparatus tested
CD—Capacity divider
CRO—Cathode-ray oscillograph

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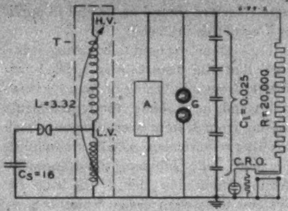

Figure 2. Circuit for switching-surge tests on high voltage power apparatus; *R* in ohms, *C* in microfarads, *L* in henries

C_s—Surge-current generator
C_1—Load capacitance
G—200-centimeter sphere gap
A—Apparatus tested
R—Resistance divider
T—2,500-kva power transformer
HV—230/$\sqrt{3}$-kv winding (230-kv insulation class)
LV—13.8-kv winding (15-kv insulation class)
L—Effective 60-cycle inductance referred to high voltage side

axiomatically in all cases even for grounded-neutral systems. For instance, on grounded-neutral systems where reduced rated arresters are applicable, power transformers in the next lower insulation class have been applied in increasing number because of the economic advantage. In the event future developments indicate possibilities of greater reduction in the insulation than now is considered good practice, then system overvoltages correspondingly will require greater attention.

Surge overvoltages inherently are higher on systems which operate with the neutral isolated and may attain five to six times the line-to-neutral crest voltage, whereas on a solidly grounded neutral the surge overvoltages seldom will exceed it $3^1/_2$ times. It is apparent that on systems which operate with the neutral isolated correspondingly more attention must be given to the insulation from a surge-overvoltage standpoint.

The foregoing considerations assume even greater importance when attention is turned to systems for operation at extra high voltages, for instance 400 kv, as have been proposed. In view of the excellent service record of present 230- and 287-kv transmission systems,[12,13] it would appear that from a consideration of lightning alone the impulse-insulation levels of these systems should be satisfactory at 345 or 400 kv. In other words, the insulation requirements from a switching-surge viewpoint will require attention at the extra high voltages. At these voltages economic advantages most certainly will dictate a grounded neutral system.

In short, there is a distinct need for fundamental data relating to the switching-surge strength of station and line insulation to supplement the present impulse data which in general may be considered reasonably complete.

Characteristics of Switching Surges

The tests of this investigation were conducted with surges of the duration and wave form given in Figures 4 and 5. The periods of these surges are in excess of 1,000 and 3,000 microseconds, respectively. Even on extensive systems, switching overvoltages seldom reach durations greater than a few thousand microseconds. Fundamentally, a longer period would mean that the system would be operating close to 60-cycle resonance, a condition rarely, if ever, attained in service. A survey of the recent contributions on system transients[14–21] indicates that periods within the range of 20 to 2,000 microsecond microseconds comprehend practically all conditions of service.* The surges selected for this investigation are adequately long and therefore, from a dielectric standpoint, correspond in effect to switching surges in the field. Furthermore, the test voltages applied to the

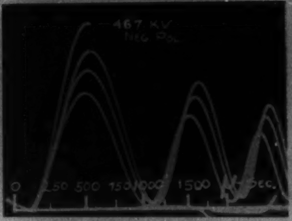

Figure 4. Typical oscillograms of switching surges applied to 46-kv bushing

equipment extended up to 900 kv crest. An overvoltage of this amplitude corresponds to nearly five times line-to-neutral crest voltage on a 230-kv system and to $3^1/_2$ times on 345 kv. In this respect, too, the wide scope of this investigation is apparent.

Apparatus Tested

The apparatus tested in this investigation are listed in Table I; some are shown in Figures 1 and 3. The list comprehends some of the more typical electric apparatus encountered in stations and on lines. For instance, switching-surge tests were conducted on outdoor apparatus bushings in the 46-, 115-, 161-, and 230-kv insulation classes. Representative of porcelain equipment are the 6-unit suspension insulator string and the 3-unit apparatus-insulator assembly. Inasmuch as air gaps are common to practically all outdoor

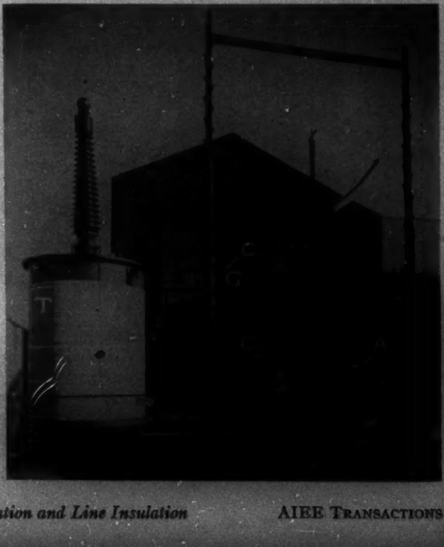

Figure 3. Partial view of switching-surge setup for testing high voltage power apparatus (refer to Figure 2 for circuit)

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* Figure 1 of reference 21 is an example of a switching surge recorded on a power system. Other typical examples are found in various technical papers (for example, AIEE and International Conference on Large Electric High-Voltage Systems) relating to faults and circuit-interruption investigations on power systems.

apparatus, standard rod gaps at 10-, 20-, and 40-inch spacings were tested. In the same manner, completely assembled power transformers in the 138- and 230-kv insulation class were the object of extensive switching surges in the course of this investigation and in its forerunner.[8] In view of the importance of oil as a criterion of the internal strength of station apparatus, tests on oil for representative electrode arrangement and spacings are reported here. Some of the tests were made on the apparatus dry (bushings, insulators, and so forth), others under artificial rain (wet test), and still others under condition of heavy dew. For comparison purposes impulse tests were made simultaneously on some of the equipment, and were omitted on others for which standard data are available. In short, it was the intent that the investigation be conducive to practical results and, to this end, apparatus as used in service was tested in nearly all instances.

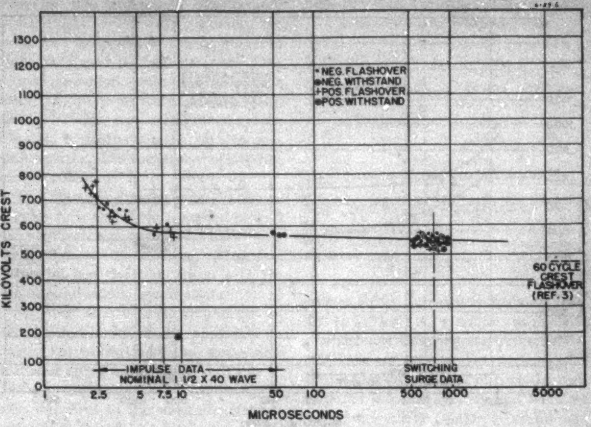

Figure 6. Switching-surge and impulse characteristics of 6-unit suspension insulator string

Relative air density = 1.00
Absolute humidity (grains per cubic foot):
Switching data, 5.1 to 8.5
Impulse data, 3.4 to 6.2

Test Methods

All switching surges in this investigation were produced by the two methods shown diagrammatically in Figures 1 and 2. The method of Figure 1 closely follows the previous investigation.[8] It consists of a generator C_g which discharges into a load capacitance C_l through an inductance *L*. The duration and wave form of the surge across the apparatus tested is fixed by the constants of the circuit, and the magnitude is controlled by the charging voltage applied to the generator. Typical oscillograms of the surges developed by this method are given in Figure 4. The maximum voltage available for test was 500 kv. Furthermore, other limitations restricted this test method to the small equipment not exceeding the 69-kv class.

Most of the tests of this investigation were conducted according to the method of Figure 2. In this method, a capacitor bank of 16 microfarads is charged to the desired voltage, and then discharged into the low voltage winding of a power transformer. The high voltage winding connects from the high voltage terminal (230-kv insulation class) to a load capacitance C_l and to a resistance divider *R*. Thus, through autotransformer action, switching surges of much higher amplitude than had been attained heretofore in the laboratory could be generated. The circuit constants were proportioned to produce surges of desired wave form and duration. Typical oscillograms of tests are shown in Figure 5.

Figure 3 gives a view of the testing equipment that was used. The capacitor banks C_g and C_l normally a part respectively of a high current and a high voltage impulse generator are located in the laboratory. The 230-kv power transformer *T* is of standard design, except that provision is not made for radiators which for the purpose of this paper are not necessary. With radiators and cooling devices the transformer would have a load capacity substantially above its present nominal rating of 2,500 kva. Therefore, electrically, the circuit elements in these tests compare favorably with major component parts encountered on large power systems. For instance, at the full charge of 100 kv the capacitance C_g has stored energy corresponding to a 30,000-kilovar capacitor

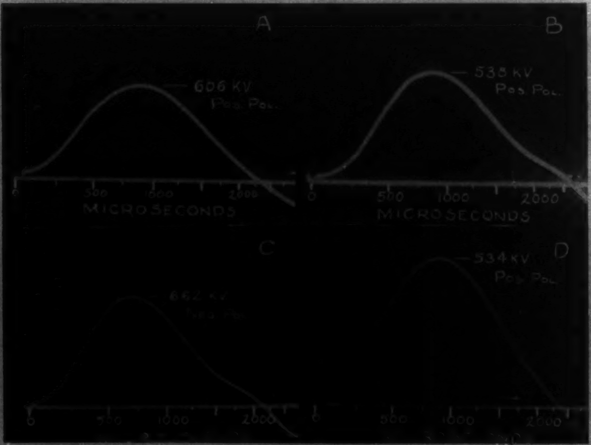

Figure 5. Oscillograms of switching surges

Surge applied to

A. 3-inch oil gap
B. 6-unit suspension insulator string
C. 115-kv bushing
D. 3-unit apparatus insulator assembly

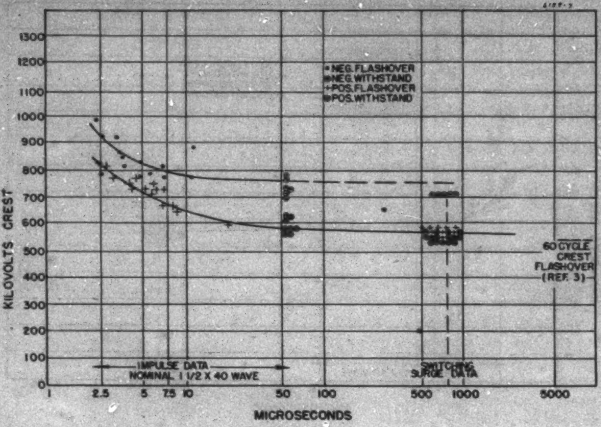

Figure 7. Switching-surge and impulse characteristics of 3-unit apparatus insulator assembly

Relative air density = 1.00
Absolute humidity (grains per cubic foot):
Switching data, 6.8
Impulse data, 7.1

bank on a 60-cycle system. The significance of these tests is borne out even more emphatically by the fact that the switching surges produced are comparable in magnitude and duration to the values recorded on power systems.

In the arrangement of Figure 2, the magnitude of the voltage developed at the apparatus tested *A* is controlled by the charging voltage applied to the capacitor bank C_s. The procedure in testing was as follows. At the start, the surge voltage would be well under the withstand value of the apparatus and then gradually would be raised, usually by 5-per-cent steps, either to the full withstand value or to critical flashover. All testing on large and small apparatus alike, both dry and wet, was outdoors in the yard adjacent to the laboratory (Figure 3). All measurement equipment is located in the laboratory. Measurements of the wave form and magnitude of the surge were effected by the recognized standard method which consists of the cathode-ray oscillograph in combination with a wire-wound resistance divider shown schematically in Figure 2. Some check tests were made between the resistance divider and a capacitance divider; these showed very close agreement in wave form between the two methods. To expedite the testing, frequently the voltage amplitude alone would be measured with a standard 200-centimeter sphere gap, using standard calibration data.[22] However, in the course of the investigation some check tests of the 200-centimeter sphere gap were made with the resistance-divider cathode-ray oscillograph method. The two methods of voltage measurement in most cases agreed within three or four per cent.

Test Data

Table I summarizes the switching-surge tests conducted in this investigation. Figures 6, 7, 8, and 9 present positive and negative switching-surge data respectively for a 6-unit suspension-insulator string, a 3-unit apparatus-insulator assembly, a standard 115-kv bushing, and and for a 1-inch oil spacing between a ball and plate. During the investigation, impulse tests were made on the apparatus using an identical setup; these results are presented in the curves for comparison purposes. In each case, the combined data serve to link the impulse and switching characteristics into a single volt-time curve.

In like manner, Tables II and III present data for 46-, 115-, 161-, and 230-kv bushings. Switching-surge data on standard rod gaps for 10-, 20-, and 40-inch spacings are reported in Table IV. No impulse tests were made on the rod gaps because data are generally available; the impulse data in the tabulation are standard reference values. The effect of dew formation and rain on switching-surge flashover of apparatus, such as in bushings and insulators, is apparent from the data in Tables II and V. Other tests of this character both on switching surge and impulse were conducted; these results are not readily conducive to tabulation and will be discussed in the following section. It will be noted that the data in these tables and figures are corrected to

Table I. Switching-Surge Tests Conducted on Station and Line Insulation

Insulation Tested	Type	Test Method	Typical Oscillograms	Polarity of Surges and Test Conditions		References and Comments
				Positive	Negative	
Bushing, 46-kv						
Without gap	Station	Figure 1	Figure 4	Dry and dew	Dry and dew	Tables II and III
With 15-inch gap	Station	Figure 1	Figure 4	Dry	Dry	Tables II and III
Bushing, 115-kv, without gap	Station	Figure 2	Figure 5	Wet and dry	Wet and dry	Tables III and V; Figure 8
Bushing, 161-kv, without gap	Station	Figure 2	Figure 5	Dry	Dry	Table III
Bushing, 230-kv, with $48^1/_4$-in. gap	Station	Figure 2	Figure 5	Dry	Dry	Table III
Suspension insulator, 6-unit string (10-in. diam × $5^3/_4$ in. spacing)	Station and line	Figure 2	Figure 5	Wet and dry	Dry	Figure 6 and Table V
Apparatus insulator, 3-unit assembly (17-in diam × $14^1/_2$ in. spacing)	Station and line	Figure 2	Figure 5	Wet and dry	Dry	Figure 7 and Table V
Standard rod gap (air), 10-, 20-, and 40-in.	General	Figure 2	Figure 5	Dry	Dry	Table IV
Oil gap, 1 to 8 in. (6-in. diam bushing ball to tank)	General	Figure 2	Figure 5			Figure 9
Power transformer						
138-kv insulation class	Station	Reference 7	Reference 7			Subjected to more than 500 surges up to 580 kv
230-kv insulation class	Station	Figure 2	Figure 5			Subjected to more than 800 surges up to 900 kv

Table II. Critical Flashover of 46-Kv Bushing to Impulse and Switching Surges

Test Condition of Bushing	Impulse, Kv* ($1^1/_2$x40 Wave)		Switching Surge, Kv* (Figure 4)	
	Positive	Negative	Positive	Negative
Bushing with 15-in. gap	260	350	240	330
Bushing without gap	285	380	290	380
Bushing without gap, dew on porcelain**			290	360

* Relative air density = 1.00 (corrected). Humidity range = 5.3 to 7.1 grains per cubic foot.

** Air terminal of bushing chilled with dry ice, resulting in heavy frost with subsequent dew formation on porcelain.

unity air density, but no attempt has been made to correct for the standard humidity reference since correction factors are still uncertain. The humidity range during the test is indicated in each test.

In addition, switching-surge tests were made in this investigation on the 230-kv 2,500-kva transformer and on a 138-kv 1,000-kva transformer as reported in the previous investigation.[8] The extent of these tests is apparent from the last column in Table I. Likewise, impulse tests up to the respective levels were made on the transformers.

Discussion of Results

With the one exception in Table II, the test data in Tables II, III, and IV and in Figures 6, 7, and 8 correspond to apparatus that is dry and in a reasonably clean condition. Compared to the full-wave impulse strength, the data invariably show little or no reduction in insulation strength to switching surges. For instance, the difference is only five per cent for either the suspension-insulator string, or the apparatus-insulator assembly, and practically none for the 46- and 115-kv bushings. Although impulse tests were not conducted on the 161- and 230-kv bushing or on the rod gaps, a comparison can be made with reference to the available impulse data on this equipment. In this case, too, little or no difference is found between the switching-surge and impulse values.

The data for the dry tests indicate good consistency throughout. As expected, the data in Figures 6 and 7 show a perceptible polarity effect for the apparatus-insulator assembly and practically none for the suspension insulator string. The small polarity effect in the 115-kv bushing data (Figure 8) apparently indicates a similar electrode arrangement at line terminal and ground for the setup. Because of the pronounced dissymmetry which exists for the 46-kv bushing between line terminal and the large ground plate (see Figure 1), the polarity effect becomes particularly marked. From the foregoing, it can be stated that the positive switching-surge strength comprises the data of greater interest since in general the strength of the insulation is lowest for this polarity.

The 60-cycle dry flashover values in Figures 6, 7, and 8 are from design tests on identical equipment or from well-recognized reference data. They are presented here for comparison with the impulse and switching-surge data.

Wet test data on switching-surges for the 115-kv bushing, the apparatus-insulator assembly, and the suspension-insulator string are reported in Table V. In these tests, the rate of water precipitation was close to the standard reference value of 0.20-inch per minute, except for the suspension-insulator string which was 0.26-inch per minute. Compared to the dry flashover, the switching-surge wet flashover values for the precipitation indicated are 20, 8, and 15 per cent lower for the bushing, apparatus-insulator assembly, and suspension-insulator string, respectively. It might be noted that the resistivity of the water in these tests was less than one-half the standard. In these rain tests, it was observed further that the strength of the insulation would recover to the dry test value immediately upon removal of the water spray, even though the porcelain and other component parts still would be dripping wet. For instance, the positive critical flashover of the bushing rose from 450 to 590 kv the instant the water was shut off. Similarly, the insulator string increased in flashover from 460 to 530 kv. An attempt was made to establish test data for precipitations under 0.20-inch per minute, but because of a breeze blowing and lack of time the tests were not disposed entirely and therefore the results can be considered useful only as a guide. They do indicate, however, that for the more usual precipitation encountered, for example 0.10 inch per minute, the switching-surge strength of outdoor apparatus is unlikely to be affected by the rain more than five to ten per cent.

The findings noted on the dripping wet insulators and bushings just reported, correlate quite well with the results of tests conducted on the 46-kv bushing to determine the effect of heavy dew forma-

Table III. Critical Flashover of Power Apparatus Bushings to Switching Surges

Rating, Kv	Gap, In.	Switching Surge, Kv*	
		Positive	Negative
46	None	290	380
46	15	240	330
115	None	590	610
161	None	750	
230	$48^1/_4$	760	**

* Relative air density = 1.00 (corrected). Humidity range = 5.3 to 7.6 grains per cubic foot.

** Maximum voltage applied to 230-kv bushing was 900 kv (negative), the limit of the switching-surge generator.

Figure 8. Switching-surge and impulse characteristics of standard 115-kv bushing

Relative air density = 1.00
Absolute humidity (grains per cubic foot):
Switching data, 5.3 to 5.5
Impulse data, 6.5 to 7.2

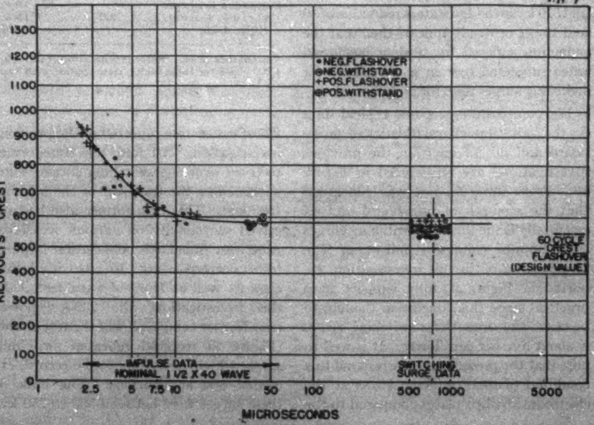

Table IV. Comparative Critical Flashover of Standard Rod Gaps to Impulse and Switching Surges*

Gap Spacing, In.	Impulse, Kv (1½x40 μsec Wave)		Switching Surge, Kv (Figure 5)	
	Positive	Negative	Positive	Negative
10	190	225	175	225
20	350		330	
40	645		655	

* Switching-surge data are from this investigation. Relative air density = 1.00 (corrected). Humidity range = 5.1 to 6.9 grains per cubic foot. Impulse values taken from American Standard C68.1.

tion, reported in Table II. The tests show that the switching-surge strength is practically unaffected by dew as compared to the strength when the bushing is dry. However, it should be noted carefully that insulation which operates in a heavily polluted atmosphere, or near the sea, or that which is subjected to heavy fogs following a long dry dusty season may be affected adversely. The effect of surface contamination on the flashover of insulation will be discussed in the next section.

For comparison purposes, wet impulse-test data also were secured on the 115-kv bushing, the apparatus-insulator assembly, and on the suspension-insulator string. Again, the bushing tests presented similar difficulties as mentioned previously because of a breeze affecting the direction of spray. Therefore, only the data on the remaining equipment are presented here. At an average water precipitation of 0.17 inch per minute, the critical full-wave positive flashover of the apparatus-insulator assembly and the suspension-insulator string is, respectively, 530 and 480 kv, compared to 590 and 570 kv in the dry state, a reduction of 10 and 15 per cent. It appears then that the reduction suffered by outdoor insulation when subjected to rain is no greater for switching surges than for impulses.

For comparison purposes typical data on the volt–time characteristics of pressboard and oil, taken from the previous investigation,[8] are reproduced in Figure 10. From these data it was concluded that the strength of pressboard and oil essentially is the same for switching surges as for impulse voltages at full wave. The test data on oil in this investigation, reported in Figure 9, fully support in a practical sense this conclusion though, to be exact, the strength for switching surges is about five per cent lower. It is well to note that the present tests correspond to a condition of more uniform stress at the electrodes (ball-to-plate) compared to the preceding tests (disk-to-plate), and in addition the tests are carried to higher voltages. The present tests therefore give support and general acceptance to the foregoing conclusion.

Though tests on oil and solid insulation provide a criterion with which to evaluate the internal strength of oil-immersed apparatus, actual tests on the complete apparatus when possible are even more conclusive in this respect. The switching-surge tests summarized at the bottom of Table I therefore have particular significance. The tests on the 138-kv power transformer were reported in the previous investigation. All told, this transformer received over 500 switching surges, some approaching the flashover of the bushing (530 kv). This transformer also had received successfully on various occasions, American Standards Association impulse tests corresponding to its insulation class as well as front-of-wave tests. In this investigation, the 230-kv power transformer being a part of the test circuit (Figure 2) received repeated switching surges which, according to the record, exceeded 800 applications. The bulk of these surges were between 500 to 750 kv, a small number from 750 to 850 kv, and a few approached 900 kv, the limit of the switching-surge generator. Previous to and during this investigation, this transformer also received successfully American Standards Association impulse tests at full 230-kv level as well as front-of-wave tests up to 1,400 kv. Since 1939 this transformer has been part of the high voltage laboratory equipment and had been used extensively in the course of prior impulse and 60-cycle investigation as well as for the purpose of excitation in commercial impulse tests.

Table V. Wet Critical Flashover of Bushing and Station Insulators to Switching Surges*

Apparatus	Switching Surge, Kv		Rate of Water Precipitation, In. Per Min
	Positive	Negative	
Bushing, 115-kv	450	495	0.23
	450		0.18, 0.18, 0.20
Apparatus insulator, 3-unit	525		0.22, 0.20, 0.18
Suspension insulator, 6-unit	460		0.25, 0.27, 0.28

* In these tests, relative air density = 1.00 (corrected) and water resistivity = 2,400 ohms per cubic inch. For corresponding dry critical flashover, refer to Figures 6, 7, and 8.

Strength of Insulation to Switching Surges

This and the previous investigation show that the switching-surge strength of insulation does not differ appreciably from the impulse strength. For example, in the directly comparative data for bushings and other high voltage insulation in air (Figures 6, 7, and 8), the switching-surge strength, in no instance, is less than 95 per cent of the impulse. Likewise, the internal strength of oil-immersed apparatus is nearly as high on switching surge as on impulse tests. In this connection, it is well to point out that stresses in the winding proper are likely to be distributed much more uniformly for switching surges. The foregoing conclusions are based on switching surges of long duration as in Figures 4 and 5. Furthermore, they apply for conditions referred to in standards as "usual." The switching-surge strength of equipment, otherwise conforming to usual conditions, is not affected by rain or dew to a greater degree than the impulse. This investigation thus establishes that the strength of station and line apparatus to switching surges in

Table VI. Comparative Data on Critical Flashover of Insulators (For Source of Data See Reference 10)

Apparatus Tested			Critical Flashover on Positive Polarity in Per Cent of Dry Flashover of 1x50 μsec Wave		
Description	50-Cycle Flashover, Kv RMS	Condition*	1x50 μsec	1x1,000 μsec	One-Half Cycle Wave of 5,500 μ Sec Duration Above One-Half Amplitude
Bushing	102	Dry	100	97	97
		Dew	98	90	80
		Rain	80	79	76
Insulator assembly, 2-unit pillar type (11-in. diam)	224	Dry	100	96	96
		Dew	92	86	84
		Rain	99	89	76
Suspension insulator string, 4-unit (10-in. diam)	227	Dry	100	99	97
		Dew	93	93	82
		Rain	85	76	72

* Insulators clean. Water precipitation (rain) = 0.12 inch per minute. Water resistivity (rain and dew) = 4,000 ohms per cubic inch. Dew applied with vaporizer.

most cases is as high as 90 to 95 per cent of the impulse strength.

Important studies on the flashover voltage of insulators in air in the range intermediate between impulse and operating frequency have been reported by W. Wanger and his associates.[10,11] The insulators tested differ from similar equipment of United States design, and the wave form of the voltages in the two investigations also differ. Consequently, it is not possible to compare the results of the foregoing contributions directly with the data in this paper. However, the data in Table VI are of interest as they provide a general comparison regarding the effect of wave form, rain, and dew on insulator flashover. The tabulation is in reasonably good agreement with the relationship established in this investigation. The research by Wanger also contributes useful data on the effect of various voltages on insulator flashover resulting from heavy contamination of the surface with soot coating, salt, and cement dust in combination with rain and dew. Under these unusual conditions, it is well-known from operating experience that the strength of insulators is reduced materially. According to his tests, in general, a pronounced reduction begins to take place for voltages of duration beyond approximately 1,000 microseconds. When unusual conditions as these prevail, the practice is to clean the insulators at frequent intervals or, in case such measures are not feasible, to provide insulation specifically to meet these conditions.

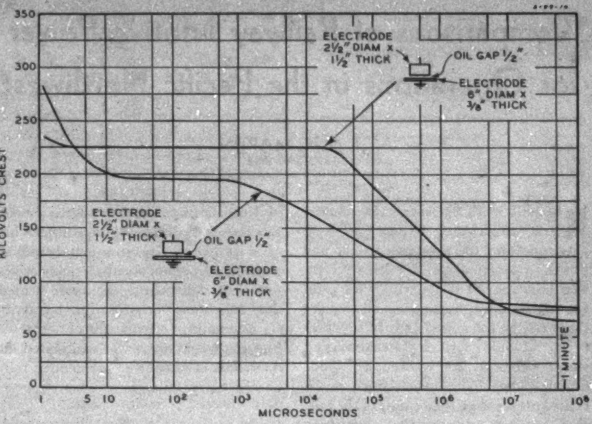

Figure 10 (right). Volt-time breakdown curves for transformer oil and oil-impregnated pressboard in oil (reference 7)

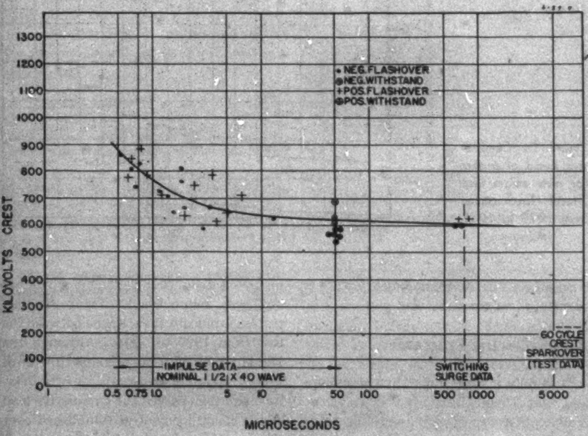

Figure 9 (below). Switching-surge and impulse characteristics of 1-inch oil gap from 6-inch-diameter bushing ball of 161-kv bushing to tank wall

As the duration of the voltage applied is extended into the range from one-tenth to one second, the characteristics of the insulation approach more and more the low frequency test values. For overvoltages of such long duration (as might develop on systems with the neutral isolated, on hydro systems, from accidental contacts between high and low voltage lines, and so forth), the performance of insulation should be assessed more with reference to the low frequency test data as well as the rating of the apparatus for continuous voltage operation.

Conclusions

1. The switching-surge strength of station and line insulation, in most cases, is 90 to 95 per cent of the corresponding impulse value.

2. For insulation in air and conforming to usual conditions, the switching-surge strength is not affected by rain or dew to a greater degree than impulse.

3. Additional studies on the effect of unusual conditions upon the strength of insulation in air appear desirable, particularly at short and medium 60-cycle durations and for sustained operation.

4. The fundamental data presented here should be conducive to a more effective utilization of insulation.

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Comparisons of Railway Motive Power for Operations in the Pacific Northwest

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RECIPROCATING steam, Diesel-electric, and electric locomotives compete today for the main-line freight and passenger assignments of Pacific Northwest railways. It is doubtful if the present struggle for dominance among the three will be influenced greatly by other types of locomotives. Reciprocating steam locomotives have occupied a position of leadership here as elsewhere in the United States since the coming of the railroad, but it is apparent that their days of unchallenged leadership are over. On the basis of physical ability to perform present and prospective tasks as well as on the basis of operating economy, steam locomotives face strong competition from Diesel-electric and electric motive power.

There have been other times when competition for steam locomotives appeared about to materialize, but each time something intervened to remove the immediate threat to their leadership. Between 30 and 40 years ago it was predicted freely that electrification was coming into recognition. Several things happened. World War I was fought and labor and materials that might have been invested in a great network of electrified railroads was used for other things. The natural advocates of electrification failed to agree among themselves on the question of what kind of a system of electrification should be encouraged. Fundamental objectives were lost sight of in the more heated than enlightened debates over the merits of alternating versus direct current. Enough uncertainty was cultivated in the minds of already steam-minded railroad officials to cause indefinite postponement of most electrification plans.

Diesel-Electric Progress

It is only a little more than ten years ago that Diesel-electric passenger locomotives first appeared in main-line service. They came with the early streamliners and to many casual travelers the words streamliner and Diesel are synonymous. It is likewise only about five years ago that the first of the so-called 5,400-horsepower 4-cab Diesel-electric main-line freight locomotives was delivered. Diesel-electric switching locomotives were in use quite a few years before either of the main-line classes, but the history of all three can be studied fairly well by going no further back than 1929. Table I is a summary of locomotive orders during the past 17 years.

Prior to the depression very few Diesel locomotives were ordered; the steam locomotive has as yet no important competition. By 1937 the Diesel was approaching equality with steam in the competition for new orders, and commencing in 1938 it assumed a leadership which has not been relinquished. Figure 1 illustrates graphically the struggle between the two from 1929 to 1946. Although the Diesel remained a poor competitor until 1940, it took but two years for it to overhaul and then exceed the cumulative lead of the steam locomotive established over the preceding decade. This is quite an

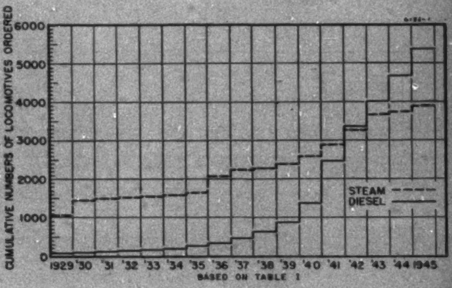

Figure 1. Cumulative trend of orders for new steam and Diesel locomotives from 1929 to 1945 inclusive

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accomplishment when it is remembered that during the greater part of these 17 years the available Diesel units were confined to switching and high-speed passenger types. Not until 1942 were road service freight units generally available. The fact is that main-line types are still relative newcomers to motive-power rosters, and the major statistical gains have been achieved by switching units. It is clear that the reciprocating steam engine has yielded to the Diesel in the field of switching service. For several years the orders for steam locomotives have included few or none of the switching type, whereas Diesel orders have run strongly to switching units. In 1945, for example, of the 148 steam locomotives ordered only 3 or about 2.0 per cent were switchers, while 478 or about 68.0 per cent of the 702 Diesels ordered were switching units.

Technical and economic factors that have enabled Diesel switching units to cause obsolescence among parallel steam types are operative in the field of main-line motive power and therefore warrant brief recital.

1. Diesel-electric units are able to motorize all their weight and hence produce relatively much greater tractive effort or hauling ability than otherwise comparable steam units.

2. With no fires to clean, ashes to dump, boilers to wash, and with less lost time for assorted other reasons, Diesels have demonstrated a higher degree of availability and reliability.

3. Diesel-electric switching units have demonstrated that they can replace steam switchers and almost invariably do a better job cheaper.

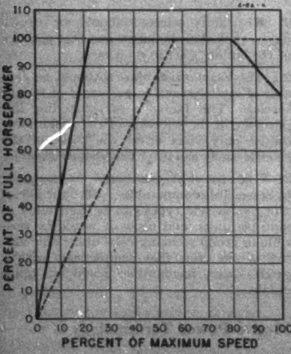


Figure 2. Percentage of full horsepower available at various percentages of maximum speeds in motor-generator and a-c series motor type locomotives

——— Motor-generator locomotives
- - - - - A-c series motor locomotives

Savings are realized chiefly in the items of fuel and running repairs. It has been established that Diesel units can cut repair costs in half and reduce fuel expense by two-thirds when they replace steam. Recent studies reported by the Committee on Railway Location and Operation of the American Railway Engineering Association under the chairmanship of L. K. Sillcox[1] suggest that savings of from one to two dollars per yard-locomotive-hour are common where Diesels replace steam locomotives. Since in yard service less total investment frequently is required for a pool of Diesel switching locomotives it can be seen that there are many situations where continued steam operations are plainly uneconomic.

Table I. Locomotives Ordered for Domestic Use, 1929–1945[7]

Year	Steam	Diesel	Electric	Total
1929	1,055	80	95	1,230
1930	382	18	21	421
1931	62	21	91	174
1932	5	7		12
1933	17	25		42
1934	72	37	76	185
1935	30	60	7	97
1936	435	77	24	536
1937	173	145	36	354
1938	36	160	29	225
1939	119	249	32	400
1940	207	492	13	712
1941	302	1,104	38	1,444
1942	363	894	12	1,269
1943	413	635		1,048
1944	74	680	3	757
1945	148	702	6	859

Note: United States Government purchases excluded for years 1942–1944.

Electrification Affected by Trend Toward Diesel

Paradoxically the trend toward Diesel-electric switching units has advanced the cause of main-line electrification. Prior to the advent of the Diesels, it was necessary to erect a great deal of trolley before a railroad could achieve full benefits from electrification. Unless electric locomotives could operate over a relatively large mileage of industrial and classification yard trackage some other kind of motive power was necessary. The capital cost of providing relatively little-used trolley facilities was excessive, and yet if both steam and electric motive power were to be required at many intermediate and all principal terminals many of the important arguments for main-line electrification were weakened. If main-line movements only were to be handled electrically, leaving yard movements to be made with steam, the opportunities for smoke abatement would be reduced seriously. Since steam and electric locomotives differ markedly in certain types of repair and service facilities many of the best opportunities for savings in operating expenses would be lost.

Table II. Principal Weights, Dimensions, and Proportions of the Steam Locomotive Used for Comparative Purposes

Class of Locomotive—4-6-6-4 Single Expansion, Articulated

Weights in working order, pounds:	
On drivers	404,000
On front truck	101,000
On trailing truck	122,000
Total engine	627,000
Tender fully loaded	437,000
Engine and tender fully loaded	1,064,000
Weights of supplies carried, pounds:	
25,000 gallons water	208,000
28 tons coal	56,000
Lengths:	
Driving wheel base	12 feet 2 inches
Total engine wheel base	60 feet 4.5 inches
Engine and tender wheel base	106 feet 8 inches
Engine and tender, coupler to coupler	119 feet
Driving wheel diameter, inches	69
Boiler and firebox details:	
Steam pressure, pounds per square inch	280
Length of combustion chamber, inches	108
Length over tube sheets, inches	240
Grate area, square feet	132.2
Direct heating surface, firebox and combustion chamber, square feet	500
Circulators (5-security), square feet	81
Total direct, square feet	581
Indirect (tubes and flues), square feet	4,214
Superheater (Elesco type E), square feet	2,162
Total evaporative and superheater, square feet	6,957
Cylinders:	
Number	4
Diameter, inches	21
Stroke, inches	32
Rated tractive force of engine at 24.1 per cent adhesion, pounds	97,350
Ratios, per cent:	
Weight on drivers to engine weight	64.4
Weight on drivers to total weight (engine and tender fully loaded)	38.0

Development of Diesel-electric switching power has improved the basis for main-line electrification. It no longer is considered proper, much less necessary, to erect trolley over any track not clearly essential to the operation of main-line trains. Operations over other trackage are within the recognized sphere of the Diesel switching unit. A Diesel-electric locomotive is related more closely to an electric locomotive than either of them is

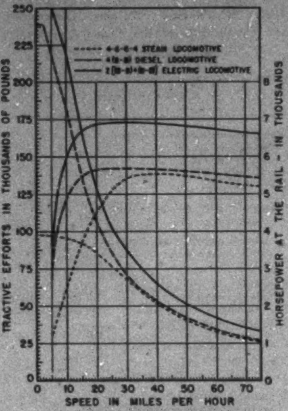


Figure 3. Tractive effort and horsepower at the rail versus speed, steam, Diesel, and electric locomotives chosen for comparison

to a steam locomotive, and joint use of many kinds of facilities and maintenance staffs is entirely practicable. A railroad company contemplating main-line electrification with a corollary use of Diesel power for yard service now can look forward to complete elimination of facilities such as coal and ash handling equipment, water service treating plants, tanks, penstocks, boiler washing facilities, turntables, and much heavy shop equipment largely useful in repairing steam locomotives. Modern Diesel and electric locomotives can utilize similar electrical maintenance facilities. Maintenance of Diesel prime movers will require specialized equipment, but the over-all shop layout is simplified and cleaned up when steam units are withdrawn. A combination of electrification and dieselization offers the next best thing to complete electrification in the reduction of smoke and fumes.

Fundamental Differences in Motive Power

Comparisons of motive power are not easy to make and will not prove satisfying or acceptable to everyone. Physical comparisons are complicated and produce differences of opinion which are not lessened when the subject is extended to include economic phases of motive power utilization.

A steam locomotive and a Diesel locomotive can be compared without too much trouble. The electric locomotive causes most of the difficulties when comparisons are attempted. This is because both steam and Diesel locomotives are so-called independent units, requiring only coal or oil to burn and track to run on to ready them for service. The electric locomotive is a dependent unit and in addition to track requires suitable central station, transmission, and trolley facilities to generate and bring electric energy to its pantographs. Except for the location of the primary source of energy, modern Diesel road locomotives and modern electric locomotives are quite similar. Both of them are in reality electric locomotives. In fact, motor-generator type electric locomotives closely resemble late model Diesel-electric road units in many important respects. One leading manufacturer is using the same type of d-c traction motors on both locomotives. The d-c generators of the two locomotives also are not too different. The chief difference in the machines lies in the means of driving these generators. Generators on Diesel locomotives are connected directly to the Diesel engines, whereas on motor-generator locomotives they are connected directly to synchronous motors which receive their energy from a relatively high voltage a-c trolley.

Basic Factors Governing Comparisons

Railroads in the Pacific Northwest are comparatively long thin lines of relatively low traffic density. A large part of all mileage is single-tracked, heavy grades are commonplace, and there is a substantial amount of horizontal curvature. In all of this region there is very little of what rightfully could be called high speed track. High speed track here is track that will permit 75.0 miles per hour with safety and comfort, and there are not too many opportunities for sustained running at that speed. Another characteristic of importance is the fact that the roads in this region are predominately used as freight carriers. This is a further reason why extremely high speeds are neither necessary nor desirable, and it is a circumstance that suggests the wisdom of choosing a kind of motive power that will deliver high tractive efforts at reasonable speeds rather than offer large horsepower outputs at speeds which are seldom attainable under actual operating conditions.

Table III. Principal Weights, Dimensions and Proportions of Diesel-Electric Locomotive Used for Comparative Purposes

Class of Locomotive—4 (*B-B*)

Weights in working order, pounds:*	
On drivers	954,000
Total locomotive fully loaded	954,000
Supplies carried and weights, pounds:	
4,800 gallons Diesel oil	36,000
3,200 gallons water	26,700
580 gallons lubricating oil	4,000
76 cubic feet sand	7,600
Lengths, feet:	
Driving wheel base	9.0
Total coupler to coupler	193.0
Diameter driving wheels, inches	40
Engine: four 16-cylinder V-type 2-cycle, rated 1,500 horsepower each	
Generator: four nominal 600-volt direct current, single bearing direct-connected to engine crankshaft	
Traction motors: 16 d-c roller bearing truck-mounted	
Gear ratio	3.53/1.00
Rated tractive force of locomotive at 25 per cent adhesion, pounds	238,500
Minimum speed (at continuous output), miles per hour	17.0
Maximum speed, miles per hour	75.0
Ratio of weight on drivers to total locomotive weight, per cent	100.0

* Electric brakes and train heating equipment included.

It is necessary to limit the scope of the comparisons to be made, and it is believed that this can be accomplished best by considering but a single locomotive from each of the classes. This imposes a considerable burden upon anyone who chooses the particular locomotives to represent each of the classes. It is obvious that the one who picks the locomotives and selects the bases of comparison can in some measure determine the outcome. Unfortunately subjective influences cannot be cancelled out entirely, and the only practical course is to strive for as much objectivity as possible.

There are a few general principles that guided the choice of locomotives for this particular comparison.

1. The units selected should be up to date—typical of the latest and best practice of the builder's art in each case.

2. The locomotives chosen should be capable of interchangeable passenger or freight service. This factor is of controlling importance in the selection of the steam locomotive only, since either the Diesel or the electric easily can meet this requirement.

3. The size of the respective units should be reasonably comparable. Both the Diesel and the electric types can be controlled in multiple units, and hence it would be possible to group so many cabs under the control of one crew that comparisons with a single steam locomotive would be quite unfair.

Motive Power Terminology

Tractive effort and horsepower at the rail are the fundamental characteristics that determine almost everything else. Tractive effort is merely the force produced by the torque of the locomotive driving wheels. It is exerted tangentially to the driving wheels and is limited by the torque and by the adhesion factor existing between wheel and rail.

In general

$$\text{Power} = \frac{\text{force} \times \text{distance}}{\text{time}}$$

or

$$\text{Power} = \text{force} \times \text{velocity}$$

If tractive effort is substituted for force and everything is expressed in proper units an expression defining rail horsepower can be obtained.

Let

V = speed in miles per hour
RHP = rail horsepower
TE = tractive effort in pounds

Then

$$RHP = TE \times V \times \frac{5{,}280}{33{,}000 \times 60}$$

$$RHP = TE \times \frac{V}{375} \qquad (1)$$

$$TE = \frac{375}{V} \times RHP \qquad (2)$$

On western mountain grades it is customary to consider only the continuous rail horsepower of locomotives having electric traction motors. The Association of American Railroads in their 1930 recommendations regarding traction motor ratings specified that the 1-hour output be determined with the motors starting "cold" (the ambient temperature). The temperature rises permitted vary with the class of insulation but are only 15 degrees centigrade higher for the 1-hour than for the continuous ratings. It is clear therefore that the hour ratings might be useful in flat country but in a mountainous region traction motors are worked at or near their continuous rating most of the time leaving little likelihood of using short-time ratings except at the start of engine runs. Even downgrade operation affords no relief since regenerative braking produces temperature rises of the same order of magnitude as does motoring. Steam locomotives present no problem of traction motor limitations, and hence for any given speed whatever rail horsepower is available will be continuously available.

Other terms used rather freely are defined in the following paragraphs.

DRAWBAR PULL

Drawbar pull is the force (either push or pull) which is exerted at the drawbar or coupler of the locomotive. It is produced by the movement of the locomotive and is equal to the tractive effort diminished by the tractive resistance of the locomotive considered as a car. Being dependent upon tractive effort it is limited also by the adhesion factor. Thus, for example, when for any reason a locomotive's drivers slip, its drawbar pull instantly and drastically is reduced. The only function of a locomotive is to create drawbar pull which can be applied to a train of cars and cause them to be accelerated into motion.

ADHESIVE WEIGHT

That part of the total weight of a locomotive which is carried on the drivers is described as adhesive. In the case of a steam locomotive it is always less than the total weight of the locomotive. In electric and Diesel locomotives, it may or may not be equal to the total weight depending upon whether or not there are unpowered axles.

ADHESION FACTOR

The adhesion factor is the ratio between the adhesive weight of a locomotive and the maximum tractive effort that can be exerted without slipping the drivers. This ratio is dependent for the most part upon the coefficient of friction between wheel and rail which is in turn largely dependent upon the condition of the rail. It usually is stated in reverse order or as the maximum percentage of the weight on drivers that can be utilized as tractive effort under varying rail conditions. This ratio varies with speed, is a maximum at starting, and drops rapidly to a value which is maintained fairly well throughout the practical speed range of a locomotive. Generally accepted values at starting are as follows:[3]

1. Clean dry well-sanded rail and rolled steel wheels—30 per cent.
2. The same conditions and cast-iron wheels—25 per cent.
3. Clean dry rail with either kind of wheels—20 per cent.
4. Wet greasy rail—15 per cent.

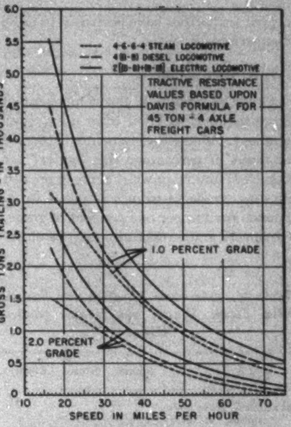


Figure 4. Gross tons trailing versus speed for 1.0- and 2.0-per-cent grades; steam, Diesel, and electric locomotives chosen for comparison

Table IV. Principal Weights, Dimensions, and Proportions of Straight Electric Motor-Generator Type Locomotive Used for Comparative Purposes

Class of Locomotive—2 [(*B-B*)+(*B-B*)]

Weights in working order, pounds:*	
On drivers	1,034,000
Total locomotive fully loaded	1,034,000
Supplies carried and weights, pounds:	
2,400 gallons water	20,000
1,000 gallons fuel oil	7,000
Lengths, feet:	
Driving wheel base	8.5
Coupler to coupler	148.0
Motor-generator sets, number	2
Traction motors, number	16
Gear ratio	3.75/1.00
Rated tractive forces of locomotive at 25 per cent adhesion, pounds	258,500
Minimum speed (at continuous output), miles per hour	17.0
Maximum speed, miles per hour	75.0
Diameter of driving wheels, inches	43
Ratio of weight on drivers to total locomotive weight, per cent	100.0

* Train heating equipment included.

Running adhesions are considerably less. Most authorities agree that 16 or 18 per cent is all that ought to be assumed in rating locomotives. Higher running adhesions are possible under ideal rail conditions and liberal sanding but should not be relied upon.

TRACTIVE RESISTANCE

Tractive resistance is the sum total of the forces which oppose motion in both the train and locomotive. It is made up of several different components which are caused respectively by

1(a). Journal resistance.
1(b). Flange resistance.
1(c). Air resistance.
2. Grade resistance.
3. Curve resistance.
4. Acceleration.

The first three are usually grouped and called net resistance on straight level track at uniform speed in still air. The latter has been measured empirically by many investigators and curves based upon the work of one of the most widely accepted authorities have been used in these comparisons.[3]

GRADE RESISTANCE

Grade resistance is the component of force required to lift a train up a rising gradient. Where the gradient is expressed in terms of percentage rise (or feet per hundred) and train weights are expressed in tons, grade resistance is equal to 20 pounds per ton per one per cent of grade. Obviously grade resistance may be either additive or subtractive depending upon whether movement is against or with the grade.

Table V. Maximum Loadings and Speeds With Steam, Diesel, and Electric Motive Power

	Steam	Diesel	Electric
Maximum gross tons trailing at 17.0 miles per hour*			
Level track	5,625**	5,625**	5,625**
1.0 per cent	3,140	4,510	5,550
1.5 per cent	2,090	3,080	3,810
2.0 per cent	1,500	2,280	2,850
Maximum speeds of freight trains, miles per hour*			
1,500 gross tons trailing*			
Level track	75.0	75.0	75.0
1.0 per cent grade	38.5	40.5	46.5
1.5 per cent grade	27.0	30.0	35.5
2.0 per cent grade	17.0	23.5	28.0
2,280 gross tons trailing*			
Level track	65.0	68.0	75.0
1.0 per cent grade	27.5	30.0	35.0
1.5 per cent grade	‡	21.5	26.0
2.0 per cent grade	‡	17.0	20.5
2,850 gross tons trailing*			
Level track	60.0	62.0	68.0
1.0 per cent grade	21.0	25.0	30.0
1.5 per cent grade	‡	18.0	21.5
2.0 per cent grade	‡	‡	17.0
Maximum speeds of passenger trains, miles per hour*			
1,200 gross tons trailing†			
Level track	75.0	75.0	75.0
1.0 per cent grade	44.0	46.0	53.0
1.5 per cent grade	33.0	35.0	41.0
2.0 per cent grade	24.0	28.0	33.0

* Resistance for 45-ton 4-axle freight cars. Speed is the minimum permissible for continuous operation at maximum rail horsepower output.

** Limited by train length to 125 45-ton cars.

† This train could be composed of 16 75-ton standard weight cars, 20 60-ton lightweight cars, or various combinations. The speeds shown for passenger trains are somewhat conservative since freight train car resistance was used in their calculation.

‡ Overloaded.

CURVE RESISTANCE

Where the gradient is constant there is an increase in resistance going around a curve which is roughly proportional to the degrees of curvature. Curve resistance also has been the subject of many empirical investigations. Western roads generally employ a figure of about 0.7 pound per ton per degree of curvature.

COMPENSATED GRADE

If mountain railroads were to have uniform gradients, operation over them would be very difficult since the element of curve resistance would introduce a persistent variable into the equation of total tractive resistance. Western railroads have found it better to plan their vertical and horizontal alignments so that, over a given section, the summation of the components of grade and curve resistance will be constant. Practically this is done by fixing a uniform average gradient of say 2.2 per cent for a section of track and then reducing this grade around all curves by an amount which for most trains nearly will balance the extra resistance imposed by the curve. If curve resistance is taken at 0.7 pound per ton per degree, since grade resistance per ton per cent is equal to 20 pounds, the equivalent grade reduction on a 6.0-degree curve would be $\frac{0.7\times6.0}{20}=0.21$.

Table VI. Evaluation of Physical Characteristics and Hauling Abilities of Steam, Diesel, and Electric Motive Power Units Selected for Comparison

	Steam	Diesel	Electric
1. Total weight of locomotive fully loaded, tons*	532	477	517
2. Total length of locomotive and tender, feet*	119	193	148
3. Rail horsepower continuously available at 17.0 miles per hour**	4,098	5,582	6,800
4. Pounds per horsepower [(line 1×2,000)/(line 3)]	260	171	152
Index, electric = 100	171	112	100
5. Feet per 1,000 horsepower [(line 2×1,000)/(line 3)]	29.0	34.6	21.8
Index, electric = 100	133	159	100
6. Maximum gross tons trailing on 1.0 per cent grade†	3,140	4,510	5,550
7. Maximum gross tons trailing on 2.0 per cent grade†	1,500	2,280	2,850
8. Hauling efficiency on 1.0 per cent grade, gross tons trailed per ton of locomotive (line 6/line 1); index, steam = 100	5.9 100	9.45 160	10.74 182
9. Hauling efficiency on 2.0 per cent grade, gross tons trailed per ton of locomotive (line 7/line 1)	2.82	4.78	5.51
Index, steam = 100	100	170	195

* From Tables II, III, and IV.

** From Figure 3.

† From Figure 4 and Table V.

The greatest advantage of employing compensated gradients is in safer, steadier operation, but it offers a further advantage to railroad students. They can forget all about curve resistance in their calculations—it has been "compensated," and the net resistance caused by both grade and curve is as though it were all concentrated in the grade. One further thing needs to be noted regarding compensated grades and that is their directional limitations. The grade can be compensated in one direction only. It is the practice to compensate the grade for up-grade movement. It necessarily follows that on down-grade movements there will be some slowing effect on curves because curve resistance is now subtractive from the component caused by grade.

RESISTANCE TO ACCELERATION

The resistances previously mentioned apply to movement without acceleration. A force is required to accelerate a mass and the amount of acceleration produced will be roughly proportional to the magnitude of the force and inversely proportional to the mass, $F=ma$.

In railroading the forces resisting acceleration are not computed easily. It is usually the practice to add all other resistances together and subtract their sum from the total tractive effort available at any given speed (it is different for each speed), and the remainder is the force available for acceleration. Thus by an integration process the time–distance relationships involved in acceleration are calculable.

MINIMUM AND MAXIMUM SPEEDS—GEARING

Electric-traction-motor locomotives require two speed limitations in connection with their ratings—a maximum and a minimum. The upper limit is absolute and is fixed by the banding of traction motor armatures. The lower limit is not absolute but depends upon the load being drawn. The series motors commonly used for traction purposes have the characteristic of increasing torque as speed drops. But as speed drops, armature current and consequently temperature rises. When series motors are turning out their continuous horsepower rating there is a certain minimum speed that must not be violated. When speed falls to the critical value and excessive amperage beings to flow, the unit is overloaded and there is no alternative but to unload the motors by reducing the trailing tonnage. Obviously, a railroad ordinarily does not dispatch trains on an experimental basis figuring that when they have stalled because of overloading they will set out excess tonnage and continue their runs. Each class of locomotive is rated; that is, it is assigned a maximum number of trailing tons that it will be permitted to depart with from a given terminal.

Locomotive Ratings

The rated tonnage of a locomotive is the maximum trailing load it can handle over its district. It is not based upon ideal conditions of weather, rail, or fuel, but upon average everyday experience with some allowance perhaps for even worse circumstances. Locomotive ratings are not equal over all districts of a railroad, but vary with what is known as the ruling grade of the particular district. The ruling grade is the maximum grade against which movement must be made save in rather exceptional cases where maximum grades are of such short duration and so located that momentum may be relied upon to assist trains over them. In this case the ruling grade will be the next heaviest grade. On all parts of a district having less than the ruling gradient a locomotive will be able to do either or some of both of two things.

1. Pull faster the same number of trailing tons it is rated for on the ruling grade.
2. Pull an increased number of tons at the same speed as on the ruling grade.

A railroad generally likes to load a train with respect to available motive power so that a reasonable mileage may be run without changing the loading. In general then the tendency will be to use the remaining power potential of a locomotive operating on less than the ruling grade in the production of more speed rather than in trying to add trailing tons. It is also clear that for this reason the determination of how much territory to include in a given locomotive district is very important. If a district is so chosen that the ruling grade, though heavy, is of short duration, and the remainder of the profile

Table VIII. Fuel and Energy Requirements and Equivalents

	Freight	Passenger	Total
Steam operations			
Pounds of coal per train	109,440	78,900	
Number of trains per day	11.4	6.0	17.4
Pounds of coal per day	1,247,616	473,400	1,721,016
Pounds of coal per year	455,379,840	172,791,000	628,170,840
Tons of coal per year	227,690	86,395	314,085
Diesel operations			
Gallons Diesel oil per train	2,476.8	1,542.0	
Number of trains per day	7.5	6.0	13.5
Gallons of Diesel oil per day	18,576.0	9,252.0	27,828.0
Gallons of Diesel oil per year	6,780,240	3,376,980	10,157,220
Electric operations*			
Purchased kilowatt hours per train	51,970	20,720	
Number of trains per day	6.0	6.0	12.0
Purchased kilowatt hours per day	311,820	124,320	436,140
Purchased kilowatt hours per year	113,814,300	45,376,800	159,191,100
Fuel and energy equivalents			
Pounds of coal per kilowatt hour	4.00	3.81	3.95
Pounds of coal per gallon of Diesel oil	67.16	51.17	61.84
Kilowatt hours per gallon of Diesel oil	16.79	13.44	15.67

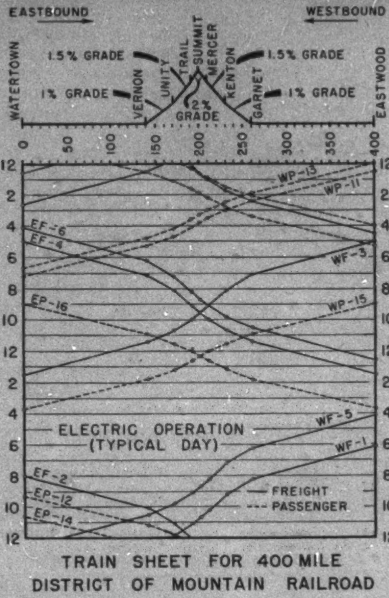
* All electric energy is measured at the low side of the 3-phase 60-cycle transformers supplying the frequency changer stations.

Table VII. Time Table Comparing Maximum Performances of Steam, Diesel, and Electric Locomotives With Tonnage Freight Trains and First-Class Passenger Trains Over a Hypothetical 400-Mile District of Mountain Railroad*

				Freight Trains						First Class Passenger Trains					
				Steam Locomotives 1,500 Gross Tons Trailing		Diesel Locomotives 2,280 Gross Tons Trailing		Electric Locomotives 2,850 Gross Tons Trailing		Steam Locomotives 1,200 Gross Tons Trailing		Diesel Locomotives 1,200 Gross Tons Trailing		Electric Locomotives 1,200 Gross Tons Trailing	
Mile-post	Station	Average Compensated Grade	Mileage Interval	Time, Hours and Minutes	Average Speed, Mph	Time, Hours and Minutes	Average Speed, Mph	Time, Hours and Minutes	Average Speed, Mph	Time, Hours and Minutes	Average Speed, Mph	Time, Hours and Minutes	Average Speed, Mph	Time, Hours and Minutes	Average Speed, Mph
0.0	Eastwood			0:00		0:00		0:00		0:00		0:00		0:00	
		0.0	140.0		75.0		68.0		68.0		75.0		75.0		75.0
140.0	Garnet			1:54		2:06		2:06		1:54		1:54		1:54	
		+1.0	30.0		38.5		30.0		30.0		44.0		46.0		53.0
170.0	Kenton			2:42		3:06		3:06		3:36		2:36		2:30	
		+1.5	21.5		27.0		21.5		21.5		33.0		33.0		41.0
191.5	Mercer			3:30		4:06		4:06		3:18		3:12		3:00	
		+2.0	8.5		17.0		17.0		17.0		24.0		28.0		33.0
200.0	Summit			4:00		4:36		4:36		3:36		3:30		3:18	
		−2.0	8.5		21.2		21.2		21.2		28.0		28.0		28.0
208.5	Trail			4:24		5:00		5:00		3:54		3:48		3:36	
		−1.5	21.5		30.7		30.7		30.7		36.0		36.0		36.0
230.0	Unity			5:06		5:42		5:42		4:30		4:24		4:12	
		−1.0	30.0		43.0		43.0		43.0		50.0		50.0		50.0
260.0	Vernon			5:48		5:24		6:24		5:06		5:00		4:48	
		0.0	140.0		75.0		68.0		68.0		75.0		75.0		75.0
400.0	Watertown			7:42		8:30		8:30		7:00		6:54		6:42	
Average speed over district					51.9		47.1		47.1		57.1		58.0		59.7
*KGTML***				812.8		1,102.8		1,346.8		692.8		670.8		686.8	
Fuel and energy requirements															
Coal, pounds				109,440						78,900					
Diesel oil, gallons						2,476.8						1,542.0			
Kilowatt hours								51,970						20,720	
Fuel and energy ratios															
Coal per *KGTML*, pounds				135						114					
Diesel oil per *KGTML* gallons						2.25						2.30			
Kilowatt hours per *KGTML*								38.6						30.2	

* Westbound movements only are shown, but since the profile is symmetrical, the eastbound movements are identical.

** *KGTML* = thousands of gross ton miles including locomotives.



is relatively easy as to gradient but is not high speed track, then the efficiency of motive power utilization will be reduced considerably. The trailing tons will be less than could be handled over most of the district aside from the small section embracing the ruling grade and the speed will not be increased sufficiently to use up the power potential of the locomotive. The working load factor of the locomotive will be low suggesting inefficient utilization.

Figure 5 (above). Train sheet for hypothetical 400-mile district of mountain railroad

Helper District

It is this general situation which gives rise to what are known as "helper" districts. The theory is to load a locomotive for some gradient less than the ruling grade, thus permitting it to take an increased number of trailing tons. When the ruling grade section is reached a "helper" locomotive is added which is capable of supplying sufficient additional drawbarpull to negotiate the hill. A helper district is nothing more than an expedient to avoid the necessity of breaking trains into smaller units for negotiation of short sections of profile. It is usually cheaper to run one train than two since although an additional engine crew may be required there is no, or at least a very minor, increase in train crew and other operating costs. Notwithstanding these favorable aspects, helper districts are nuisances and are resorted to only where they cannot be avoided.

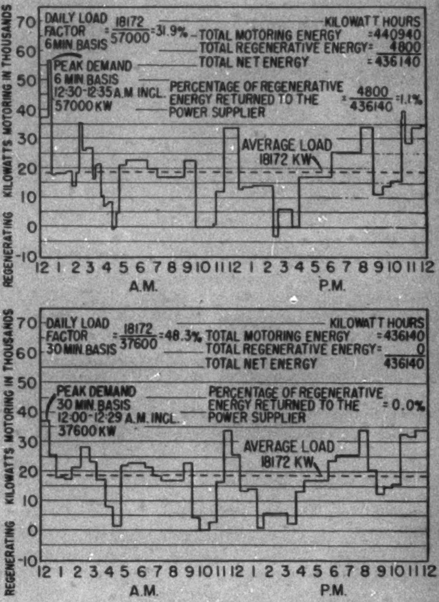
Motive Power Chosen for Comparison

Characteristic curves and basic performance data used in these comparisons were assembled from a number of sources and are believed to be factual and fair. For obvious reasons exact identification of the locomotives used has been withheld.

Steam Locomotive

Table II is a brief description of the steam locomotive selected for this comparison. Known in motive power circles by its wheel arrangement as a 4-6-6-4 it is of the single expansion articulated type. It is a 4-cylinder engine, in reality two separate engines served by one boiler and under the control of one engine crew. It bears the closest resemblance to multiple unit control available where electric motors are not used for traction purposes. With a driving wheel diameter of 69 inches a reasonable compromise is had between the objectives of high tractive effort demanded in freight service and high maximum speeds requisite to passenger operations. The top speed of the engine should be about 75.0 miles per hour and it appears to be about as satisfactory a universal passenger-freight type as is available for use in this part of the country.

Figure 6 (below). Substation load charts for typical maximum day of electrified operation shown in Figure 5



Diesel Locomotive

There are at least four manufacturers building so-called 6,000-horsepower Diesel-electric locomotives at the present time. The one chosen here is of the 4-cab type with each cab having a B-B wheel arrangement. Table III lists the principal facts regarding its weight, size, and so forth. The gearing was selected with the purpose of limiting its top speed to approximately match the top speed of the steam locomotive chosen or 75.0 miles per hour. This fixes the minimum speed under continuous output conditions at about 17.0 miles per hour. It should be noted that if the steam locomotive chosen actually will not run this fast, the use of the 3.53 to 1.00 gearing chosen will penalize the Diesel slightly. For example, if the top speed could be taken as 70.0 miles per hour a ratio of about 3.81/1.00 could be used which would permit the minimum speed under continuous output to be dropped back to 15.5 miles per hour. This might not seem important, but it actually would afford about a 10-per cent increase in the continuous tractive effort of the Diesel and an equal augmentation of its hauling ability. The assumption therefore is that all of these locomotives will be able to attain top speeds of 75.0 miles per hour. Oil fired steam generators have been included for passenger train heating purposes.

Table IX. Motive Power Availability and Utilization

	Steam Operation			Diesel Operation			Electric Operation		
	Freight	Passenger	Total	Freight	Passenger	Total	Freight	Passenger	Total
1. Road time required by one train, hours	7.7	7.0		8.5	6.9		8.5	6.7	
2. Number of trains per day	11.4	6.0	17.4	7.5	6.0	13.5	6.0	6.0	12.0
3. Total road hours per day	87.78	42.00	129.78	63.75	41.40	105.15	51.00	40.20	91.20
4. Total road hours per year	32,040	15,330	47,370	23,269	15,111	38,380	18,615	14,673	33,288
5. Assumed utilization of availability, per cent*			50			50			50
6. Total hours locomotives should be available for assignment			94,740			76,760			66,576
7. Number of locomotives			15			11			9
8. Total potential locomotive hours per year			131,400			96,360			78,840
9. Derived availability factor, per cent (line 6/line 8)			72.1			79.7			84.4

* This factor permits calculation on a uniform basis for all three classes of motive power, of the difference between actual time spent in trains on the road, and the total time the pool of locomotives is available for work assignments. It accounts for all differential time except that required for running and classified repairs.

ELECTRIC LOCOMOTIVE

A somewhat more detailed explanation of the choice made in the case of the electric locomotive seems appropriate. It is obvious that the selection of the electric locomotive also implies selection of the type of system to be employed. What kind of electricity will be placed on the catenary for the pantographs to pick up? It also should be obvious that it is entirely beyond the scope of the present paper to delve into the ramifications of a choice of system. The purpose here was to compare some electric locomotive with the best available steam and Diesel motive power under parallel operating conditions. Naturally, it was desired to select an electric locomotive equal to the competition. It would be pointless to pick one either physically or economically deficient.

Regarding the choice of system, the primary criterion was the ability to deliver at least 15,000 kw to a point on the trolley system. If enough substations could be built there would be no particular bar to delivering such an amount of power, but, as even the casual observer knows, railway power supply facilities cost a great deal and there are few opportunities to be lavish with substations if over-all economics mean anything. Under the circumstances the use of a low voltage trolley system appears to be impractical. To keep power supply facilities, including both the elements which normally are railroad-owned and those normally owned by the power supplier, at a minimum it seems essential to keep the trolley potential up to at least 12 kv. From the immediate practical standpoint this definitely suggests the necessity of choosing a 12-kv 25-cycle trolley system.

The choice of a frequency of 25 cycles requires explanation. There are many reasons why the frequency should be lower than 60 cycles, but few if any of them can be cited to justify the choice of 25 cycles. The Swiss and Germans appear to have made a wiser choice with their $16^2/_3$-cycle equipment. American practice probably would be better had a simple 3/1 or 4/1 ratio to the basic 60 cycles been established a long time ago. There are now too many precedents and too much installed equipment to be upset to suggest 15 or 20 cycles, and even a switch to 24 cycles, which would permit substantial improvements in frequency changers, appears difficult. Be this as it may, the necessity of lowering frequency from 60 cycles is virtually unchallenged. It is necessary because of at least two things.

1. The excessive reactance of 60 cycles.
2. The single phase unbalance that would be sustained by a power supplier attempting to feed 60 cycles to a railroad for direct application to the trolley. Changing frequency permits the assembly of what otherwise would be single-phase low-power-factor loads into a 3-phase unity-power-factor load.

With the choice of the trolley system fixed at 12 kv, 25 cycles, it is necessary to choose between two types of motive power.

1. Locomotives using single phase a-c motors of the series commutator type for traction.
2. Locomotives using d-c series motors for traction with d-c generators directly connected to synchronous motors supplied from the a-c trolley.

Fortunately the choice between these two is comparatively easy in this part of the country. The a-c traction motor locomotive has certain characteristics that make it a relatively inferior performer in a region where heavy gradients prevail and where freight forms a large part of total traffic. In such circum-

Table X. Summaries of Train Crew and Locomotive Maintenance Expenses

	Steam	Diesel	Electric
Traffic statistics			
Number of freight trains per day	11.4	7.5	6.0
Freight train miles per day	45,600	30,000	24,000
Freight train miles per year	1,664,000	1,095,000	876,000
Number of passenger trains per day	6.0	6.0	6.0
Passenger train miles per day	24,000	24,000	24,000
Passenger train miles per year	876,000	876,000	876,000
Total locomotive miles per year (Line 3 plus line 6)	2,540,400	1,971,000	1,752,000
Train crew costs per year*			
Freight train miles at $.80 per mile	$1,313,200	$ 876,000	$ 700,800
Passenger train miles at $.33 per mile	$ 289,080	$ 289,080	$ 289,080
Total train crew costs	$1,620,280	$1,165,080	$ 989,880
Locomotive maintenance costs per year**			
Steam locomotive repairs at $.55 per locomotive mile	$1,397,220		
Diesel locomotive repairs at $.55 per locomotive mile		$1,084,050	
Electric locomotive repairs at $.32 per locomotive mile			$ 560,640

* Assumed on basis of comparable wage rates prevailing in the Pacific Northwest. Freight crews consist of engineman, fireman, conductor, and three brakemen. Passenger crews consist of engineman, fireman, conductor, and two brakemen.

** Locomotive maintenance costs assumed on the basis of the best available experience that is believed credible.

stances the prime requirement of motive power is the ability to provide high tractive effort at relatively low speeds. This in turn means that the horsepower characteristic of a locomotive must be such that full horsepower is available in the speed range where it can be used effectively. Figure 2, which follows recent analyses of F. E. Wynne of the Westinghouse Electric Corporation, illustrates the nature of the horsepower characteristics of the motor-generator and a-c commutator type locomotives.[4]

It should be noted that full horsepower is available in the case of the motor-generator locomotive in the range from 22 to 80 per cent of full speed, whereas in the case of the a-c commutator locomotive the full horsepower range extends from 56 to 100 per cent of full speed. At 22 per cent of full speed the a-c commutator machine achieves only about 40 per cent of its full horsepower. Since the tractive effort of a locomotive at a particular speed varies directly with its rail horsepower, it is evident that at 22 per cent of full speed a motor-generator locomotive would have 2.5 times the tractive effort of an a-c commutator locomotive of equal nominal horsepower. Such a deficiency in tractive effort means a corresponding decrease in the number of trailing tons which the locomotive could be loaded with for operation over the ruling grade. So long as the major premise of the desirability of the relatively low speed operation of freight trains in a region such as the Pacific Northwest is granted, it is hard to see that any choice other than the motor-generator locomotive is warranted. There are compelling reasons for the supplier of power to prefer the motor-generator locomotive, but it is believed unnecessary to labor this point when there appears to be no logical reason for a railroad to have a contrary preference.

Table XI. Assumed Capital Cost of Facilities Affected by Type of Motive Power

400-Route-Mile District of Mountain Railroad

Affected Items	Steam	Diesel	Electric
Electrification fixed properties			
480 track miles of trolley and single phase transmission construction			$ 5,760,000
Frequency changer apparatus and trackside substations			3,650,000
Communications and signal changes and other contingencies			590,000
Total fixed properties (electrification)			$10,000,000
Motive power			
15 (4-6-6-4) steam locomotives at $250,000 each	$ 3,750,000		
11 Diesels at $600,000 (61,400 continuous rail horsepower at $107.49 per horsepower)		$6,600,000	
9 motor-generator electric locomotives at $525,000 each (61,200 continuous rail horsepower at $77.21 per horsepower)			4,725,000
Total affected capital costs	$3,750,000	$6,600,000	$14,725,000
Incremental investment over that required for steam operation		$2,850,000	$10,975,000

Table XII. Assumed Annual Operating Expenses of Facilities Affected by Type of Motive Power

400-Mile District of Mountain Railroad—Case 1

Affected Accounts	Steam	Diesel	Electric
Operating expenses			
Maintenance of way and structures, electrification fixed properties			$125,000
Maintenance of equipment, locomotive repairs*	$1,397,220	$1,084,050	560,640
Transportation rail lines, train crew expenses*	1,620,280	1,165,080	989,880
Fixed charges			
Depreciation			
Electrification fixed properties, $33^{1}/_{3}$-year life—3.0 per cent			300,000
Steam locomotive, 25-year life—4.0 per cent	150,000		
Diesel locomotive, $16^{2}/_{3}$-year life—6.0 per cent		396,000	
Electric locomotive, 20-year life—5.0 per cent			236,250
Interest, total affected investment—5.0 per cent	187,500	330,000	736,250
Total affected expenses except for fuel and energy	$3,355,000	$2,975,130	$2,948,020
Fuel and energy expenses**			
314,085 tons of coal at $2.00 per ton	628,170		
10,157,220 gallons of Diesel oil at 6.0 cents per gallon plus 20.0 per cent for engine lubricating oil		731,320	
159,191,100 kilowatt-hours at 4.0 mills per kilowatt hour			636,764
Total affected expenses	$3,983,170	$3,706,450	$3,584,784
Annual savings over steam operation		$ 276,720	$ 398,386

* From Table X.

** Quantities from Table VIII.

Table IV includes the principal dimensions of the motor-generator type electric locomotive selected. The important points to be noted are

1. The complete motorization of all weight.
2. The gearing of 3.75/1.00 which on 42-inch wheels gives about the same maximum and continuous speeds as the Diesel, that is, 75.0 and 17.0 miles per hour respectively.
3. The provision of oil fired train heating equipment.

Horsepower and Tractive Effort Comparisons

Figure 3 is the starting point for analyses. Included are characteristic rail horsepower and tractive effort curves. Attention is called to the point at which substantially full horsepower becomes available in the different locomotives. It will be noted that in the case of both the Diesel and the electric locomotives this occurs at about 17 miles per hour but is not reached until 30 miles per hour in the case of the steam. At first this may not seem a serious defect in the makeup of the steam locomotive, but it probably will become gradually apparent that it is organic. Even if there were some assured means of shifting the characteristic so that full horsepower were to be available earlier, it probably would be found that the adhesion would not be sufficient to permit full utilization of the additional tractive effort. As it is the steam locomotive chosen for comparison is credited with about 90,000 pounds tractive effort at 17.0 miles per hour. Since the weight on drivers is but 404,000 pounds this would be equivalent to a working adhesion of 22.2 per cent, and it scarcely can be presumed that anything more would be possible even with clean dry rail and liberal sanding.

Hauling Efficiency of Locomotives

Probably the most important single criterion for judging these locomotives is their ratings in terms of the trailing loads each of them could handle on various ruling grades. These ratings have to be based upon assumed average conditions of tractive resistance, and accordingly the Davis formula for 45-ton 4-axle freight cars has been used. Drawbar pulls for various grades were computed and loading curves prepared. Space permits the inclusion in Figure 4 of curves for the 1.0- and 2.0-per-cent grades only. The curves show gross tons trailing versus speed where the trailing load is composed of 45-ton 4-axle freight cars, but can be taken as applicable to other trailing loads averaging similar tractive resistance. Table V contains a summary of freight and passenger loadings and speeds for different grades. Table VI contains a summary of the more important data that facilitate appraisal of the physical performance of these locomotives. With a weight efficiency of 260 pounds per continuous rail horsepower at 17.0 miles per hour the steam locomotive is 71 per cent heavier than the electric locomotive which weighs but 152 pounds per continuous rail horsepower. In the length efficiency comparisons the Diesel locomotive shows to less advantage requiring 59 per cent more length of cab per 1,000 continuous rail horsepower.

It is the hauling efficiency comparisons that are the epitome of most of the observations that can be made of the physical abilities of these locomotives. It should give pause for thought that on a 2.0-per-cent grade the steam locomotive will haul only 2.82 tons for each one ton of its own weight. It is seen that the Diesel locomotive will handle 4.78 tons and the electric 5.51 tons per each ton of their respective weights. In addition to the inefficiency included in this comparison, steam locomotives spend sizeable portions of their useful lives engaged in the nonrevenue task of hauling their own fuel back and forth to various fueling stations. If this were measured easily it would impair still further the hauling comparisons with electric motive power which requires no allowance for such nonrevenue trailing loads.

Economic Generalizations Hazardous

As stated earlier economic comparisons are difficult. Extreme variations in the physical conditions encountered make generalizations risky. Each railroad presents a somewhat different situation and often different divisions of the same railroad pose completely different operating problems. Actual case studies of the economics of motive power utilization require time and money and must be applied to specific sections of railroad. In view of space requirements, it is obviously impossible to analyze adequately economic factors surrounding the use of the locomotives described in the previous sections. It seems worthwhile to present some comparison however inadequate and hypothetical it may be in order to indicate the nature and general magnitude of the factors that affect the economics of motive power utilization at the present time and to suggest how changes in these factors could affect future comparisons.

Table XIII. Effect of Changes in Price of Fuel and Energy Upon Annual Expenses Shown in Table XII

	Steam	Diesel	Electric
Total affected expenses except for fuel and energy	$3,355,000	$2,975,130	$2,948,020
Case 2			
314,085 tons of coal at $2.00 per ton	628,170		
10,157,220 gallons of Diesel oil at 7.0 cents per gallon, plus 20.0 per cent for engine lubricating oil		853,206	
159,191,100 kilowatt hours at 4.0 mills per kilowatt hour			636,764
Total affected expenses	$3,983,170	$3,828,336	$3,584,784
Annual savings over steam operation		$ 154,834	$ 398,386
Case 3			
314,085 tons of coal at $2.00 per ton	628,170		
10,157,220 gallons of Diesel oil at 7.42 cents per gallon, plus 20.0 per cent for engine lubricating oil		904,399	
159,191,100 kilowatt hours at 4.0 mills per kilowatt hour			636,764
Total affected expenses	$3,983,170	$3,879,529	$3,584,784
Annual savings over steam operation		$ 103,641	$ 398,386
Case 4			
314,085 tons coal at $2.50 per ton	785,162		
10,157,220 gallons of Diesel oil at 9.00 cents per gallon, plus 20.0 per cent for engine lubricating oil		1,096,980	
159,191,100 kilowatt hours at 4.0 mills per kilowatt hour			636,764
Total affected expenses	$4,140,162	$4,072,110	$3, 584,784
Annual savings over steam operation		$ 68,052	$ 555,378

Hypothetical Operating Section

The method employed here had to be hypothetical and greatly oversimplified. A theoretical 400-mile operating district of a mountain railroad was devised. To save time a greatly simplified and symmetrical profile was assumed with all gradients combined into four classes—0.0, 1.0, 1.5, and 2.0 per cent. Contrary to actuality, grades were assumed to be compensated equally in both directions. A sketch of the profile appears in Figure 5.

A uniform daily traffic was assumed as follows:

1. Freight, 8,550 gross tons trailing in each direction.
2. Passenger, 3,600 gross tons trailing in each direction.

This traffic was assumed to be handled at maximum performance by each kind of locomotive. No allowances were made for acceleration time, station stops, meets with other trains, in short no delays of any kind; just maximum performance from terminal to terminal was assumed. Also no helper service was assumed. Each locomotive takes its rated tonnage (fixed by the 2.0-per-cent grade) and uses up any excess horsepower available on the lesser grades in additional speed.

Table VII is a time table illustrating maximum performance of freight and passenger trains. A rather curious fact revealed here is that the tractive effort characteristics of the steam locomotive are such that when it is loaded to its maximum rated tonnage for the 2.0-per-cent or ruling grade it can run faster with that load on any lesser grade than the Diesel or electric locomotives can run with their rated tonnages. Care must be exercised in interpreting this fact. It does not mean that the steam engine can run faster with the same trailing load. This is revealed clearly in the average speeds attained by the passenger trains. It does mean that the steam locomotive in freight service can trail 1,500 gross tons over these 400 miles at an average speed of 51.9 miles per hour, whereas the Diesel and electric locomotives are able to handle their rated tonnages at but 47.1 miles per hour. The trailing loads handled by the Diesel and electric locomotives, however, are 2,280 and 2,850 tons respectively. Tables VIII, IX, and X are included to summarize essential data necessary in the calculation of operating expenses.

The principal comment that is required here relates to the rather remarkable savings in maintenance that are evident in the case of the electric locomotive. It seems to be true that on an equated rail horsepower–mileage basis, an electric locomotive of the type selected here can be maintained for from 40 to 50 per cent of the corresponding cost of a Diesel of the type selected. This can be explained quite well on the basis of the work necessary. The electrical and running gear maintenance of the two will be nearly identical. Maintenance of the Diesel engine accounts for the chief difference. It can be seen that the difference in crew costs is purely a function of the relative hauling ability of the three types. Use of electric locomotives reduces the number of train miles that have to be run to a minimum that is nearly half that required by steam locomotives, and only about 80 per cent of that necessary for Diesel locomotives.

Capital and Operating Costs Compared

Table XI shows the assumed capital cost of those facilities which are affected by the type of motive power. Obviously these are not the entire capital costs of building such a section of railroad, but only the costs of the items which would vary with the different types of motive power. Considerable simplification will be noted in these assumed data. Investments other than for motive power have been eliminated from the steam and Diesel figures. Fixed properties therefore appear only in the case of the electric. This is not actually true, but it is an assumption that works no particular hardship on the steam and Diesel locomotives. The prices that have been assumed like all prices in troubled times are subject to uncertainties, but the comparison between types of locomotives is believed to be realistic and fair.

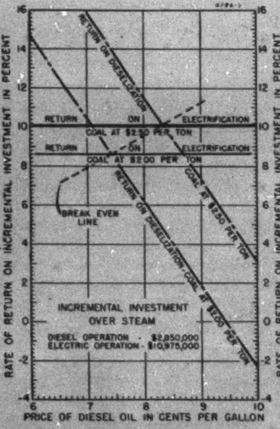


Figure 7. Effect of changes in the price of Diesel oil and coal upon the rate of return from incremental investments made for dieselization and electrification

The affected operating expenses as of the present time are presented in Table XII. Here, as in the case of the capital costs, there is room for differences of opinion. The objective has been to use the best data and experience available that seemed to be free from bias. Table XII reveals that on the basis of present prices annual savings of $276,720 are possible where Diesels are substituted for steam locomotives at an incremental cost of $2,850,000 and annual savings of $398,396 are possible by electrification at an incremental cost of $10,975,000. With the 5.0-per-cent interest previously charged this is equivalent to returns of 14.7 and 8.6 per cent respectively on the incremental investments made for dieselization and electrification respectively. If this were all that could be said for the two alternatives it would appear that a case had been made for dieselization, for on this basis about a 6.0-per-cent greater return could be earned on money spent for modernization.

Effect of Diesel Oil Price Changes

That this would be a dangerous conclusion to reach hurriedly is evident upon examination of Table XIII where an attempt has been made to show how the annual savings of the different plans of operation are affected by changes in the price of Diesel oil and coal. For example, note that in case 3 where the price of Diesel oil has been assumed at 7.42 cents per gallon, an increase of not quite 25 per cent, the annual savings on Diesel operation are reduced to $103,641. This reduces the return on the incremental investment in dieselization to 8.6 per cent, exactly equal to that obtainable from electrification. Whether it should be assumed that Diesel oil will increase in price without assuming a corresponding increase in the price of coal can be debated. There are nearly as many opinions available on the future price of both commodities as there are authorities to be quoted. There does seem to be some evidence to suggest that the long term price trends of the two are different. The supply of coal though limited is fairly well known, while the supply of oil is both limited and unknown. It can be argued that both commodities are likely to increase in price with respect to their present bases but that the factor of increasing demand is more likely to produce a disproportionate increase in the case of Diesel oil. Demands for Diesel oil and all other petroleum products are increasing by leaps and bounds while the bituminous coal industry is organizing to combat what they view as the possibility of gradually decreasing demands.

Electrification Offers Hedge Against Inflation

Figure 7 has been prepared to indicate the effect of price changes in both Diesel oil and coal upon the relative economy of dieselization and electrification. It is believed that a substantial case for electrification can be inferred from the relationships shown. The reasoning probably would be as follows:

1. The present attractiveness of Diesel-electric locomotives as replacements for steam locomotives has both a physical and an economic foundation.
2. Physically, the electric locomotive has no equal—there is nothing the Diesel locomotive can do to outclass the steam locomotive that the electric unit cannot improve upon.
3. Measured economically, dieselization has considerable present margin over steam.
4. Measured economically, electrification also offers a margin over steam although the returns on incremental investments are not so large as offered by the Diesel on the basis of present prices.
5. The crux of the contest between dieselization and electrification lies in the question of which of the two is more likely to maintain or increase its present margin over steam, and this depends almost entirely upon the future price of respective energy sources.
6. The price of electric energy over the next decade or two may not decrease markedly, but it is hardly likely to rise. This largely is based on the fact that the supply can and very probably will keep pace with the demand.
7. There is no corresponding assurance of the maintenance of present prices for Diesel oil, and most authorities apparently anticipate gradual increases.
8. If the price of coal does not increase beyond present levels (in the view of many a strictly possible result of decreasing demand), an increase in the price of Diesel oil of 25 per cent approximately would equalize the return from investments in electrification and dieselization.
9. Should coal rise about 25 per cent in price a rise of 40 per cent in Diesel oil prices again would equalize dieselization and electrification.
10. In view of the foregoing, electrification appears to offer railroad mangements the most likely hedge against inflation of operating costs. By electrifying main lines and dieselizing yard operations, it appears feasible for them to achieve some assurance of reasonable costs during the next 20 years.

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Specific Engineering Problems in Rural Electrification and Electroagriculture

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FELLOW AIEE

A DISTINCTION is made between "rural electrification" and "electroagriculture," the former being the art and the business of bringing electric service to the rural area and to the individual consumer in that area, the latter being the art and the business of using electric power for farming.

In both of these fields there are specific problems for the electrical engineer, some old, some new, some partly solved, others as yet unsolved, and still others which have not been definitely stated yet.

There are collateral problems between the two fields. One of the problems is the case of industrial loads supplied from rural lines. Offhand it would seem that a sharp line of demarcation between industrial load on the farm and that in urban areas would be important only for the purpose of statistics and economics but not for engineering purposes. However, occasionally specific engineering problems arise in connection with industrial loads on the farm. Even though very few such specific problems have been crystallized as yet, indications are that many new ones will arise as the farm load grows. Therefore a few specific items will be referred to.

While specific solved or partly solved problems in both fields will be presented, the main purpose is to point to problems yet to be solved, to others yet to be stated, and to stimulate and challenge the electrical engineer, particularly the young electrical engineer, to tackle them.

The sum and substance of all engineering problems in both fields can be stated thus: continuity of service on the farm is more important than in the city. This applies both to the continuity of the power supply and the continuity of operation of the individual piece of equipment. An outage of a few hours in the city may be a mere inconvenience. One can go to the movies, eat in a restaurant, or relax in the dark. But when a farmer uses electric milking, electric water pumping, or electric cooling or freezing equipment and the power is off for a few hours, he may experience a serious financial loss. This continuity of service however, must be rendered at a low cost. Furthermore, when a computation in large power systems is completed and the problem is known, the necessary diversity of circuit breakers, relays, regulators, and other devices generally can be selected from a catalog to solve the problem, or new relays can be designed. But in the case of the rural primary circuits, there is no such diversity; there is only the fuse, the circuit breaker with a series coil, and possibly a thermal tripping device. Thus the engineering problem of the rural circuit is in some respects more difficult to solve than that of the 460,000-volt line.

Shall a power system study the load requirements and provide a supply system to satisfy these loads, or shall it promote loads which will help build up the system? The answer is that it should do both and should not waste time discussing the question of whether the chicken or the egg was first. For practical purposes the load conditions will be discussed here first, followed by a discussion of power supply problems. What kind of specific loads are there on the farm? A list of possible uses of electrical service on the farm, prepared by J. P. Schaenzer, is presented in the appendix. It is no doubt incomplete, but it is sufficient to indicate the great diversity.

It is evident that we are dealing here with a new industry, the electroagriculture industry. A new industry calls for new engineering treatment. Only a few years ago few items of electric equipment for specific use on the farm were manufactured for the market by quantity production methods. American manufacturers thought that there was no market for such items. Competent hard-working men in agricultural colleges developed some fine designs, but there were no manufacturers willing to produce them. The best they could do was to send farmers the designs and specifications for homemade equipment. How many farmers can read blue prints, particularly when the drawings are not prepared by experienced draftsmen? How many farmers have the time for it? It became evident that if the American farm was to be electrified, it would have to be done by American manufacturers. But there was such a great diversity of ideas among agricultural experts and agricultural colleges that manufacturers hesitated to invest in tooling up for quantity production of any one design. It may be recorded that the American manufacturing industry finally did create this new industry during the war years, the most difficult imaginable time for such a creation. How did it come about? A few cases will serve best as a general demonstration.

A few years ago, few manufacturers were on the market with combination quick-freeze and zero-storage units suitable for the farm and at a price which the farmer could afford to pay. A brief survey was made among farmers. Based on this survey, a functional specification was prepared and distributed among manufacturers. Recently, several hundred manufacturers announced such models.

Manufacturers found it difficult to embark upon quantity production of pig brooders because practically every state agricultural college had a different design. An investigation was made among agricultural colleges and full co-operation was furnished by many professors. A careful study of all the diversity of designs resulted in one basic design for a pig brooder which met practically all the requirements. A report was prepared and made available to manufacturers. As a result, several manufacturers now advertise models at low prices, because they are manufactured by quantity production methods.

During the war the electric welder became a very important tool on the farm for the repair of damaged equipment because the farmer was unable to get

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either new equipment or repair parts. No data were available to advise the manufacturers of the type of welders which could be supplied properly from rural power lines. In many instances the welder loads caused distribution transformers to burn out, disturbances of voltages, and similar nuisances. An investigation was made on which was based a specification prepared in co-operation with manufacturers and this specification eventually was accepted by the National Electrical Manufacturers' Association (publication 45-105, September 1945). Now the manufacturer can proceed with quantity production and the farmer can get a welder which will not disturb the power system.

A survey was made among colleges in search of new developments which might be ready for manufacturers to develop for quantity production. The report was distributed among manufacturers and several items now are being studied by them. Here one of the difficulties encountered is the policy of some schools not to permit members of the faculty to secure patents on inventions. It is done with the best intention. They feel that a devleopment made in a public institution should be available to everyone. However, this very policy prevents many a good idea from being put to general use. No invention is ever ready for the market. Expensive development usually is called for. A manufacturer cannot be expected to spend great sums on such development unless he has some exclusive patent protection. This applies also to a so-called public patent on which everyone could secure a free license. Some institutions, such as Purdue University, have a very sound policy in this respect. Other schools should give this problem the consideration which it deserves.

Numerous items are now in the development stage but the new electroagricultural industry already has arrived and is here to stay. It offers a challenge and an opportunity to the electrical engineer.

In an industry such as this, which is still partly in *statu nascendi*, it would be premature and inadvisable to establish rigid standards which might interfere with the creative imagination of the engineer and inventor or with creative competition which is needed for progress. However, some basic standards for performance as well as size and quality of material are needed to make it possible for the buyer to judge quality. Commercial test procedures also are needed for determining the service the buyer may expect an item to render. It is suggested that a special AIEE committee be set up for this purpose. Co-operation with committees of other bodies should not be too difficult. But most of the pending problems seem to be of an electrical engineering nature.

Attention may be called to a few specific items for which there is immediate need. Two hundred years ago sanitation was not as important as it is today, as is evidenced from the requirements of a French boarding school, then patronized by the French nobility. In its catalog to parents of prospective female pupils it includes the following revealing information:

> Pupils are entitled to have one set of underclothing, one pair of stockings and two handkerchiefs per month. Towels: Pupils, one every week; nuns, one every two weeks. Foot baths: Pupils, one a month; nuns, only by special authorization of the superior. Complete baths: Three a year (May, June and July). Pupils unable to take their baths on the appointed day must wait until the following month.

Today, no one will dispute that sanitation is one of the most important requirements of human beings if they are to live on a level higher than that of pigs.

The most important need is perhaps for a small electrically driven water pump, simple and low priced, to make it possible to introduce a minimum amount of sanitation even in the poor farm shack.

A small milk pasteurizer for the farmer who has only a few cows is needed badly. The present difficulty is that in most places a curve-drawing thermometer is required which costs more than the whole pasteurizer would. The problem is either to develop a cheap temperature recorder or an electronic pasteurizer which does not require temperature recording and on which it would be possible to secure approval upon presentation of competent test reports.

No adequate data are available as yet which could be used as a basis for a functional specification for a competent electric egg cooler. Research now is being carried on, and it is hoped that it soon will be possible to furnish manufacturers the necessary fundamental information for the design and quantity production of such a cooler.

The problem of control of off-peak water heating on the farm has not been solved yet. For various reasons, clock control has not become popular. What is needed is a device which will disconnect the water heater upon the expiration of a definite time after the total load on the service transformer has reached a prescribed limit, and to reconnect the water heater as soon as other loads have been reduced. This would permit control regardless of the time of day or night. The principal difficulty in solving this problem is the fact that the device would have to be both simple and inexpensive to be suited for farm use. It is hoped that the development work now carried on by some manufacturers will result soon in a competent device.

Off-peak house heating has been tried, but in some cases it turned out to be peak-load. This definitely calls for further study. In many parts of the country the heat pump eventually may become the best method of heating and cooling farm houses and other farm buildings.

Another item for which there is a definite demand is a competent, automatic, electric high-pressure cooker.

Fire hazard on the farm is far more serious than in the city. Many farmers would like to eliminate overhead secondary distribution lines between buildings, because of the hazards of lightning, icicles, trees falling on the conductors, and roosting chickens or other birds. For single-phase circuits it is conceded generally that an underground cable made up of one insulated conductor and one parallel bare conductor would offer the best lightning protection. Manufacturers are now busy at work on the development of an economical conductor of this nature, and it is hoped they soon will be able to place one on the market which will meet the approval of the authorities. This cable should be ratproof and gopherproof. It also should be frostproof, to make it unnecessary to dig below frost line.

Perhaps the field of freezing and cooling offers greater opportunities than other items for invention, improvement, and development. It is important to repeat that while centuries have been devoted to the engineering of controlling temperatures upwards, the engineering of controlling temperatures downwards is of comparatively recent origin.

According to the United States Department of Agriculture Circular 423, January 1937, an estimated total of 60,000,000 rats on the farm cost the American farmer about $65,000,000 a year. It is indeed surprising that with this enormous amount of money involved, to say nothing of the danger to human health, American engineers have not tackled the problem yet. At least two types of electric rat catchers are known to exist and are reported to have given satisfactory results. American manufacturers should concentrate on the study of this problem and attempt to develop electric rat catchers which could be manufactured by quantity production methods and made available to the electrified farm at reasonable prices.

The study of the possibility of a small storage-battery-driven tractor has been delayed by the fact that some new lightweight batteries developed during the war are still military secrets. As soon as such batteries are available, the study should be continued. The convenience of electrical control certainly promises considerable advantages. Storage-battery charging would be excellent night load.

The application of high frequency to various purposes on the farm has not been studied sufficiently yet; neither has that of ordinary light or infrared and ultraviolet radiations.

At present, freezing, cooling, heating, lighting, water pumping, and similar electrical applications on the farm are mere adaptations of developments made for city use. Engineers should begin to place themselves in a state of mind of never having seen any of these applications, crystallize the specific problem of farm use, and then proceed to develop equipment for the specific purpose. Good results would be inevitable. Numerous specific problems no doubt soon will be crystallized in relation to the electric motor and the motor control equipment on the farm. The problem of power factor control, limitation of starting current and similar specific problems are now ready for investigation in a similar manner as the welding problem was investigated. One item, the need for which has been known for some time, is a simple and inexpensive means for stopping a motor automatically when the load is finished, as for instance for feed grinding and sawing wood, where time clocks are not suitable. One or more such devices are already on the market. But more designs for quantity production are needed.

Electroagriculture offers a challenge and an opportunity to the electrical engineer. A great deal has been done by basic research, and many results of this research are available for the engineer to pick up for industrializing. It is surprising to find how little engineering has been applied to the problem of food production, food distribution, and food preservation. This art is only in the process of becoming engineering. On the whole our food production still depends on the climate and the weather as it did many centuries ago. Freezing and air transportation are in the process of bringing about drastic changes. As freezing and high-speed air transportation become economical, it no longer will be necessary to raise everything everywhere. Fresh fruits and vegetables may be raised only in places where they can be raised with the least effort and delivered fresh to large consumption centers in a matter of hours. It is the job of the engineer to make this economical. But this is not enough. It is the ultimate job of the engineer eventually to make food independent of climate and independent of weather. It is not expected that such a revolution will be brought about in a day or a year, and it would be premature to suggest at this time how this will be brought about. But when hundreds or thousands of engineers begin to think about it seriously and continue thinking about it for 25, 50, or even 100 years, they eventually will solve the problem as they have solved many more difficult problems in the past. But the time is definitely at hand to begin engineering thinking.

It is evident that the farm load and demand diversity are, and will continue to be, of such uncertain nature as to make circuit computation difficult. But this is the time to start the study of the characteristics of this diversity and to continue it. The present cost of the numerous demand meters which would be required makes such a study practically impossible. It is necessary to resort to statistical interpretation, and very little has been done even by statisticizing. The great accuracy of the present meters, which were developed for different purposes would not be required. If a simple, cheap, recording ammeter were to be brought out, there would be a quantity market for it.

All of these features and many others must be given consideration in the planning of rural electrification (supplying power for rural areas). A layman, or even the average engineer, when observing a typical rural circuit from the window of a railroad train, will be surprised to hear that the simple little pole with a pair of small insulators, carrying a pair of conductors and occasionally a small transformer with its accessories, offers any engineering problem at all. It is therefore of some importance to show that there are engineering problems, that many of these problems have not been solved yet, and that others have not been stated. The crystallization of these problems and their solution are essential if the farmer is to have eventually the degree of continuity of service which the character of the load calls for or will call for in the future when more items of electric equipment for farming are made available. In fact, the speed with which many new uses of electric service will be introduced on the farm will depend to a great extent on the speed with which many engineering problems will be stated and solved.

Most of the primary circuits in rural areas are single-phase multigrounded 60-cycle 7,200-volts phase-to-ground. There is a main conductor and a neutral conductor. The primary and secondary windings of each distribution transformer are grounded together to the neutral, and a ground wire generally is carried down the pole to a ground rod at each distribution transformer. Grounds also are installed frequently on intermediate poles. The secondary windings are carried to the house or other building, and there usually is located another ground connection. Thus, a multiplicity of parallel grounds exist along the line, primary and secondary. Anyone who ever tries to trace the various paths that the current could take loses himself in what appears to be a hopeless maze. This is one of the most complex circuits in electrical engineering. The question as to why this specific circuit was selected instead of a delta circuit would be beside the point at this time, because there are already hundreds of thousands of miles of such circuits and the time has definitely arrived to establish a mathematical foundation for them. After consulting many outstanding authorities, to the best of the author's present knowledge, the general differential equation for the multigrounded single-phase circuit has not been set up yet. After it eventually is set up, it no doubt will be of a series character with numerous variables, and very likely it will not be possible to integrate it generally. It very likely will become necessary to assume some variables as constants for different purposes and establish a different integration for each purpose. The mathematical study is now under way, and the indications are that the basic differential equation soon will be available. Until it is available, problems of short circuit, voltage regulation, hazards brought about by voltage gradients on the surface of the soil, and similar problems cannot be solved to the full satisfaction of electrical engineering.

What goes on in the soil in the proximity of such a circuit? Very few test data or analytical data are available. The answer will not be at hand until many more tests are made and interpreted in the light of the basic formula yet to be set up. It is known that the circuit in the ground does not flow in the direction of the line from ground rod to ground rod. Tests show that the voltage distribution in the soil is nearly uniform in each direction from each ground rod. It is evident that soil capacitance will have to be given consideration in addition to soil resistance. Observations made so far indicate that the voltage gradient on the surface of the soil can be modified considerably with multiple grounds of various designs. But the available data are inadequate as a foundation for definite designs in definite cases. It is evident that a great amount of testing superimposed on mathematical analysis will be required before a definite engineering foundation is available for the solution of many related problems.

This naturally leads to the question of the ground return line, a single-phase circuit made up of one wire, the earth forming the return. On one hand there are statements about great hazards, and on the other hand the question is raised why such circuits are not used more extensively. Circuits of this nature have been used in various places, and to the best of the author's present knowledge have not resulted in accidents.

Two types of hazards are mentioned. The first is the possibility of a break in the conductor which runs down the pole to the ground. In this case there would be practically full voltage across the break even though two or more such ground conductors generally are carried down the pole. The protective covering of the conductors would have to be broken to result in this hazard. The second is the possibility of dangerous voltages on the surface of the soil. It is evident from the foregoing remarks that no definite statements of a quantitative nature can be made at this time which would serve as a guide for those who are considering the use of such circuits. Neither is there sufficient evidence to prove or disprove that there are real hazards, what their degrees are, and how to overcome them. Those who use such circuits have to exercise caution after a thorough familiarity with soil conditions and load conditions in each case. The National Electric Safety Code states in relation to such circuits:

> In urban districts supply circuits shall not be designed to use the ground normally as the sole conductor for any part of the circuit. It is recommended that such use be avoided in rural districts.

It is to be noted that in urban areas it "shall not be used" but in rural areas it is "not recommended." However, a responsibility is involved in recommending it because in case of a jury trial for criminal negligence relating to an accident, the recommendation of the National Electric Safety Code no doubt would carry considerable weight.

Analytically the ground return circuit is a special case of the general multigrounded circuit. Once the formula for the general circuit is available, the ground return circuit can be developed by making the neutral impedance equal infinity.

It is evident from this discussion that the least that can be done is to provide the best possible grounds everywhere and to inspect and test the grounds periodically, first to make sure that they are intact, and second to accumulate as much data as possible on the electrical characteristics of the specific soil. However, our present methods of testing grounds are cumbersome, and the testing equipment is too expensive. Here too the great accuracy for which the various instruments were designed is not required. With lower accuracy it should be possible to develop simple low-cost instruments for quantity production, so that each farmer who has a worth-while electrical installation could himself periodically test not only all his electric circuit grounds, but also all his lightning rod grounds. Simpler methods of attaching the testing instruments to the circuits also are required. There will be very little testing if the present procedure is to be continued, which makes it necessary in each case to dig up the soil at each ground rod, disconnect the ground lead from the rod, connect the instruments to the rod, disconnect it after the test, reconnect the ground lead to the rod, and then refill the soil. What specific engineering problems are implied in the little distribution transformer used on rural circuits? Surprising as it may seem, there are numerous complex problems. It has not been possible yet and will not be possible until the previously referred to recording ammeters are available in quantity to determine the best range of sizes of distribution transformers. The sizes now used are those originally determined for urban purposes. The weight is important because of the handling difficulties in farm areas. Numerous accessories have been added to the little transformer, such as lightning arresters, gaps, and cutouts, which make it resemble a Christmas tree. A great deal of competent design study still is called for before a reasonably satisfactory arrangement will be had.

A number of small transformers in a rural area with unsatisfactory $I \cdot T$ products may result in inductive interference, which may call for considerable expenditures in testing to determine the location of the cause of the interference and to design remedial measures altogether out of proportion to the cost of the transformers. It is essential that manufacturers are made to realize that it is more important for a rural transformer to have a satisfactory $I \cdot T$ product than it is for an urban transformer. This also applies as well to small generators in isolated areas. The cost of the remedy is less than the cost of the investigation. It is gratifying to report that most manufacturers appreciate this problem and are doing their best to reduce harmonic troubles.

One of the most important features is the need for low core losses. An urban transformer may have a diversity of loads with a reasonably good over-all load factor. A rural transformer generally has only one consumer. The copper loss load factor may be as low as 15 or 20 percent, whereas the core loss load factor is always 100 per cent. There are cases where the billed kilowatt-hours are less than the core loss kilowatt-hours which are not billed. The Kelvin law, in any one of its forms, could be set up to determine the economy of the ratio of core losses to copper losses. But the problem can be stated simply without any formula. There are three criteria.

1. The lower the fixed charges, the more investment is justified on the saving of core losses and replacing them by a corresponding amount of copper losses.
2. The higher the cost of energy is on the primary winding side, the more can be invested on the reduction of core losses.
3. The lower the load factor is on the individual transformers, the more can be invested on the reduction of core losses.

It is realized that the reduction of core losses is not a simple problem for the transformer designer if the over-all efficiency is not to be reduced, if the resulting exciting current is not to bring about troublesome harmonics, and if the voltage regulation is not to be made worse. Nevertheless, manufacturers advise that there is hope for a considerable reduction in core losses in the near future. To indicate the great strides already made in improving rural transformer design, it is only necessary to report what one large manufacturer accomplished in a short time. The weight of the $1^1/_2$-kva transformer was reduced from 275 pounds in 1936 to 200 pounds in 1937, 145 pounds in 1943, and 120 pounds in 1945. Its core losses were reduced from 24 watts to 22 watts. The core losses of the 3-kva transformer were reduced during that same time from 36 watts to 30 watts and those of the 5-kva unit from 51 watts to 48 watts. This is progress indeed. But it is not enough. Considering the fact that the cost of core losses on rural distribution transformers amounts to several million dollars per annum, further reductions definitely are called for. It is reasonable to assume that they will be forthcoming.

It was mentioned before that the absence of a diversity of relays and electrically operated circuit breakers makes circuit control in rural areas difficult. In addition, the small circuit breaker available for sectionalizing primary rural circuits has to be automatic reclosing. And yet, the performance of this little circuit breaker is nothing short of being miraculous. S. R. Faris, manager of Illinois Rural Electric Company, Winchester, Ill., reports:

> During the heavy rains and floods of the spring of 1944 a landslide occurred on a bluff parallel to our line at a point about three miles west of the oil circuit breaker. Small trees of from three to six inches in diameter were carried into our line and the oil circuit breaker went to work. The landslide occurred about eight o'clock one morning and our members reported that our lights had been coming and going from that time until four the next afternoon when we finally discovered the condition.
>
> We discovered the condition during a patrol of the line along the river. The land on which the line was built was almost all about eight feet under water. During this patrol the linemen stopped at the oil circuit breaker installation and could hear it operating once or twice every five to ten minutes. No reports of trouble had been received by our members on this line.
>
> Upon arriving at the location of the landslide, our patrolmen found that the neutral wire was broken and that quite a few of the smaller trees had been burned off and cleared the phase wire. One tree, approximately four inches in diameter, had also been burned off and cleared, and another tree about six inches in diameter was up against the phase wire and it was about half burned in two. The power was still on, however, and it was necessary to kill the line before making repairs.
>
> Upon reading the counter on the circuit breaker, we determined that it had operated approximately 93 times during this fault. After repairs were made and the trouble cleared up, the writer visited several members receiving service from the section of line controlled by this circuit breaker and questioned them as to why they hadn't reported the condition. The answers were all the same—they didn't think they should report an outage as long as the power kept coming and going and they felt sure it would clear up soon.

This is quite an accomplishment for a little circuit breaker.

The same system which installed 60 circuit breakers in 1942 reports:

> Average number of operations per year per circuit breaker 12.12 with an average of 0.25 lockouts per month. The total operating expense saved by each circuit breaker per year is estimated as over $26.00. This is based on straight time pay, even though actually about 50 per cent of the outages had to be taken care of by overtime at time and a half. The cost of fuse replacements is offset by the cost of maintaining the circuit breaker.

However, even this investment is too high for minor primary tap lines. Two manufacturers therefore now are bringing out still smaller circuit breakers at considerably lower cost because this small circuit breaker requires practically no rupturing capacity. One manufacturer, for instance, furnishes the following description:

> The sectionalizer is designed so that its contacts do not open during the time that a short circuit exists on the system. The actual clearing of line faults and interrupting of short-circuit current is performed by a backup recloser. The sectionalizer operates to disconnect the branch line only after one, two, or three operations of the backup recloser have failed to clear the line fault.
>
> When a permanent line fault occurs beyond a sectionalizer which is backed up by a recloser, the short-circuit current flowing through the sectionalizer will prepare it for an opening operation. The short-circuit current which also flows through the recloser will cause it to operate in an attempt to clear the fault. During the time which the recloser is open, the sectionalizer moves forward one position toward the opening of the contacts. The second operation of the recloser again prepares the sectionalizer for an opening operation and during the time that the recloser is open the sectionalizer again moves forward another position toward opening. After the third operation of the recloser, the contacts of the sectionalizer are opened during the time that the recloser is open. In this way the faulted branch line is disconnected and when the recloser energizes the system after the third operation, the fault has been removed, allowing service to be continued to the balance of the system.
>
> If the line fault is of a temporary nature, it will be cleared by the first or second operation of the recloser, and service to the entire system will be restored. Following this, the recloser and the sectionalizer will reset themselves automatically for another complete cycle of operations.
>
> The sectionalizer may be adjusted to open on a permanent line fault after one, two, or three operations of the backup recloser. Inasmuch as the sectionalizer opens its contacts only during the no voltage period while the backup recloser is open, the sectionalizer is not required to interrupt short-circuit current. Consequently, there is no appreciable wear of the sectionalizer contacts or contamination of the insulating oil.

Even for the primary and secondary sides of the little distribution transformer manufacturers have been at work developing automatic devices to replace fuses. The type of transformer with internal primary fuse and internal secondary circuit breaker has become very popular. One manufacturer recently placed on the market a "fuseless cutout." This device is said to produce a dead short circuit to ground on the primary side when a heavy overload occurs on the secondary side or within the transformer, resulting perhaps in a heavy arc on the primary side. The main reclosing circuit breaker then trips. As soon as this happens the arc is extinguished. When the main reclosing circuit breaker recloses, the arc is out; the individual faulty transformer is disconnected; the reclosing circuit breaker remains closed; and the whole system, except for one consumer, again has service.

One reason against the latter two types is that every time there is a fault on the tap line or on the individual transformer, the main sectionalizing circuit breaker will open for a short period and many farmers will have to reset their clocks. But a farmer would much rather have to reset a clock than be without service for a considerable period.

Anyone who is familiar with the history of switchgear knows that no item of electrical equipment has been undergoing such a continuous process of change and improvement as has the circuit breaker. No circuit breaker design ever is finished. The need for major and minor improvements continually is revealed by continuous co-operation of manufacturer and operating man. No one therefore needs to be unduly disturbed if operating experience calls and will continue to call for improvements in the design of the rural sectionalizing circuit breaker which already meets unprecedented demands. It is certain that as the need for improvement is revealed, manufacturers will introduce them.

When the load on a line has been disconnected for any appreciable length of time and the line is re-energized, a large current, several times the normal value, flows. This current decreases with time but does not return to the normal value for an appreciable period sometimes running into minutes. This re-energizing phenomenon has been termed inrush current and is caused by the current inrush into resistance type loads, and more particularly by the starting requirements of direct connected motors.

The early types of reclosers used on rural circuits had a very short opening time delay. Consequently, in many cases when the recloser was closed back in on the load for re-energization purposes, the inrush current retripped the recloser.

In many cases the recloser had to be by-passed temporarily in order to pick up the load.

Some later designs have greater opening time delay, overcoming pickup difficulty to some extent. However, the exact shape or characteristic of the inrush current as it varies with time is not yet sufficiently known. Moreover, the rapid increase in the use of motors on the farm promises to make the problem much more serious.

Tests now are being carried on to determine the shape of the inrush current time curve and to predict the ultimate shape with the load growth. Preliminary tests on one section of rural line indicate an inrush current of less than 200 per cent of the normal peak after an outage of a few seconds. After an outage of one hour the inrush current was 400 per cent of the normal peak. But in all cases the inrush currents decayed to roughly one-third of their initial values in the first two seconds. It is hoped that results of further tests will be available soon.

Small rural generating plants run fewer machines at light load than at full load. The full load current at peak, therefore, may be more than the fault current during off-peak hours. The power factor of the load current is not vastly different from that of the fault current. Inexpensive devices are needed that will distinguish between load and fault currents, and which could be applied to pole-mounted reclosers.

In some cases of transformer failures the line sectionalizing devices may operate to lock out the lines before the transformer fuses blow. The transformer then may be very difficult to find. There seems to be a vast difference in characteristics of internal fuses furnished by different manufacturers. There is a need for standardization of internal fuses and development of recloser and fuse characteristics to co-ordinate more properly.

Voltage regulation problems on rural circuits are not vastly different from those on any circuit. They are only more aggravating, first because the rural circuits extend comparatively great distances from the source, and second because line investment must be kept to a minimum.

Original estimates of load on rural systems were difficult because very little experience was available as to rural power consumption. Later experience on rural systems has shown that the original estimates were too low, and a great deal of engineering is called for on many systems to bring the voltage regulation within acceptable limits. Now that the farmer is learning how to use electric power for production, experience shows that load estimates need to be revised sharply upward.

As already mentioned, the solution of the problem of voltage regulation is first of all waiting for the setting up of the basic circuit formula. Inexpensive low-capacity regulators with small steps are needed for regulating long branch lines. The rapid changes in load characteristics which may be expected in the near future would seem to make it imperative for manufacturers to explore the possibility of inexpensive secondary regulators. Perhaps some device eventually will be incorporated in the distribution transformer.

The rapid growth of motor power on rural systems indicates an eventual need for low cost capacitor installations to switch on and off automatically with load. Series capacitors no doubt would solve many a problem, but a competent system of protection for such capacitors still remains to be developed.

The problem of fault locating is as old as the electric circuit itself. Aggravating as a long delay in locating a fault may be in urban areas, it is even more aggravating in rural areas. And again the problem is aggravated by the imperative demand to do the job at low cost.

Rural distribution systems extend into remote areas which are frequently inadequately served with telephone facilities or have no telephone service at all. Outages in such areas may be prolonged because of lack of communication with the power maintenance office. There are cases on record in which the maintenance office was notified of outages by postal card and letter. Needless to say, such loads as milking machines, coolers, chick brooders, and water pumps make prolonged outages intolerable.

A carrier current outage recorder was manufactured experimentally and was undergoing field tests at the outbreak of the war. Further work on this equipment was suspended for the duration of the war. It consisted of a small low-cost carrier current transmitter located at each important sectionalizing point and an associated receiver located at the maintenance office. When an outage occurred a coded impulse from the appropriate transmitter was sent to the receiver, which activated a punch type paper strip recorder and time stamp. Thus, the location and time of the outage were recorded immediately. Provisions were made for another operation of the recorder when service was restored so that a complete record of the outage could be maintained.

After the end of the war the study was resumed. Several schemes which offer the possibility of being simpler and more reliable than the original one are being considered now. One of these schemes may make use of the radar pulse technique for detecting circuit irregularities and discontinuities. Another scheme may make use of the change in circuit impedance when an outage occurs.

It appears today that some competent means for reporting and recording outages automatically will be solved in the not too distant future.

Mention already has been made of the fact that the main problem of inductive co-ordination in rural areas is the cost of investigation. With the great amount of research and development over many years and the unusually high degree of co-operation by power system engineers and communications engineers, there is as yet no cut and dried method either for locating the source of trouble or for designing the remedy. Each case is still practically an individual research job, and each remedy still has to be practically a tailor-made job. Special equipment is required for testing, and it takes a good engineer a considerable time as an assistant on actual cases before he can take charge of investigating a case on his own and designing the remedy. This applies equally to telephone interference, telegraph interference, and railway signal interference. A complete visualization of the simultaneous functioning of an interconnected power system is called for, superimposed on a visualization of the functioning of the communications system, in addition to a familiarity with the functioning of each part of each system. The ability for such a visualization cannot be acquired in a day. It is utterly impossible and no doubt will continue to be impossible to do such a job from a text book, even if a good text book were available, which is not yet the case. Neither could the best special college course prepare an engineer to be able to assume the responsibility without first having served as an assistant on several actual cases.

With the exception of radio, full co-operation prevails between the engineers of power systems and those of communication systems, so that neither side is called upon to carry alone the burden of the investigation and the design and installation of the remedy. Power systems and communication systems both find it advisable to have specially trained engineers available. These engineers on both sides generally communicate with each other and tackle the job jointly. It is suggested respectfully that every young electrical engineer who makes power his career make every effort to find an opportunity to participate in at least two or three inductive co-ordination investigations from the first report of the trouble, through the tests, the preparation of the report, and the installation and testing out of the remedy. It will help him acquire a visualization of the functioning of a power system. Not only do the engineers on both sides co-operate after a case of interference actually is reported, they also consult with each other when a power system or a communication system is being designed and thus prevent in advance as much as possible the appearance of interference. Research laboratories and manufacturers of power and communication equipment likewise render their services and co-operation.

In the field of radio such co-operation has not yet been established, and this applies not only to rural systems but to all power systems. When a case of radio interference is reported, the whole burden of locating the source of trouble and designing and installing the remedy has to be carried by the power system. It is hoped that it will be possible soon to establish between the engineers of power systems and those engaged in broadcasting, radio research, and radio manufacturing the same degree of co-operation that now prevails in other fields of communication.

There is now a great need for a simple, practical, and inexpensive radio interference locating instrument, specifically for rural areas. It is hoped that radio research institutions soon will bring out not just one such instrument, but a diversity of instruments. Here, as in most fields, competition will result in progress. There is definitely a quantity market for such an instrument.

In addition to the previously referred to possible use of carrier for the automatic recording of outages, carrier recently has been tried out on some rural power systems for communication purposes for power system maintenance. Here too the difference between carrier on high voltage lines and carrier on rural lines is principally one of permissible cost. The developments have not gone far enough for a competent analysis on the AIEE platform.

This applies also to 2-way radio as an auxiliary to maintain the rural power system. Some interesting and promising developments are taking place now, but they do not yet lend themselves to a competent AIEE presentation. Possibly, leased radio channels furnished by telephone systems will be the best answer, just as leased telephone wires were found by many power systems to offer advantages over private telephone circuits on the transmission towers or on separate poles.

All of this relates only to communication problems for the power system itself and not to the problem of communication for the farmer. Particularly in the case of carrier there has been some confusion of the two, and it is important to establish the distinction. Recently an unprecedented demand for telephone service developed among farmers throughout the country. This probably was brought about partly by the enormous increase in the use of mechanical and electric equipment on the farm. The challenge was taken up and is being attacked rigorously from three angles.

1. There is a great activity in the extension of direct wire communication to farms.

2. Intensive studies are being made and experiments carried on in the joint wire use of power poles for telephone and telephone poles for power.

3. Intensive experiments are being carried on in the joint use of carrier telephone on power lines to render telephone service to farmers who have electric power service.

The latter two developments are not yet ready for a competent AIEE presentation. It is sufficient to mention that numerous intricate engineering problems, some apparently picayune but actually very difficult and annoying, are involved in joint wire use and joint carrier use. In connection with carrier it is important to mention that there is a great difference between the incidental use of a carrier phone by a lineman to talk to his headquarters and the continuous use of telephone service by a domestic or commercial telephone subscriber. Several years of research at great cost have been devoted to the latter problem, both in the laboratory and in the field, before the first experimental instruments were installed in farm homes. Here again, the cost is of extreme importance. One of the bottlenecks of cost was the coupling capacitor. For a long time the cost appeared to be prohibitive, until a progressive American manufacturer developed a capacitor for a 7,200-volt line, with a dry impulse flashover of about 200 kv and at a cost of around $10. It is reasonable to say that without this development carrier telephones in quantity would not be an immediate prospect for the American farmer. They are definitely so now.

This is about all which is considered practical to present within the scope of a single paper. But it is hoped that it will serve the purpose of bringing about a realization in the electrical engineering profession that the humble rural circuit offers important engineering problems worthy of the serious consideration of the electrical engineer. It is hoped that it will result in a much broader realization of the utter necessity of co-operation between system design engineers, operating engineers, manufacturing engineers, and research engineers.

Appendix. Applications of Electricity for the Farm

Farm Home

Air conditioner
Air sterilizer
Alarm
  Burglar
  Fire
Amplifier
Aquarium heater
Attic fan
Bell
Bell ringing transformer
Biscuit baker
Blanket
Bottle
  Sterilizer
  Warmer
Bread mixer
Broiler
Bun warmer
Can opener
Casserole
Cauterizer
Chafing dish
Chimes
Cigar lighter
Clippers
Clock
  Household
  Time
Coal stoker
Coffee
  Maker
  Mill
Communication system
Corn popper
Cream whipper
Croup kettle
Curling iron
Dehydrator
Dishwasher
Door opener
Double boiler
Doughnut kettle
Dumb-waiter
Egg
  Beater
  Cooker
  Poacher
Evaporative cooler
Fan
Floor
  Polisher
  Sander
  Waxer
Flour and cereal mill
Flour sifter
Fly trap
Food
  Chopper
  Mixer
  Warmer
Freezer
  Cabinet
  Walk-in
Fruit and vegetable juicer
Fruit corer and peeler
Frying pan
Furnace fan
  Circulating
  Forced draft
Games
Garbage disposer
Grill
Hair drier
Hand drier
Heating pad
Hedge clipper
Hot plate
Humidifier
Ice cream freezer
Ice maker
Insect
  Trap
  Eradicator
Iron
Ironer
Laundry drier
Lawn
  Edger
  Mower
Lights
  Building
  Decorative
  Photoflood
  Yard
Massager
Meat
  Cooker
  Grinder
Milk shaker
Motion picture projector
Necktie presser
Oil furnace
Organ
Oven
Oven (infrared)
Paint sprayer
Pants presser
Percolator
Phonograph
Photoenlarger
Player piano
Pottery kiln

- Pump jack
- Radiation
  - Infrared
  - Ultraviolet
- Radio
- Range
- Razor
- Refrigerator
  - Household
  - Walk-in
- Roaster
- Room heater
- Sausage
  - Grinder
  - Stuffer
- Scissors
- Screen
- Sewing machine
- Shoe polisher
- Siren
- Slicer—food
- Space heater
- Steam iron
- Steam radiator
- Straw flower drier
- Sump pump
- Teakettle
- Televisor
- Toaster
- Tooth brush
- Toys
- Vacuum cleaner
- Vaporizer
- Ventilator
- Vibrator
- Waffle iron
- Washing machine
- Water
  - Chlorinator
  - Cooler
  - Heater
  - Pail
  - Softener
- Water system
  - Deep well
  - Jet
  - Shallow well
- Wiener roaster
- Woodburner—decorative

Farm Equipment

Crop Processing

- Baler
- Blower
  - Grain
  - Roughage
- Bulb cooker
- Caster bean huller
- Clover huller
- Cider mill
- Conveyer
- Corn
  - Cracker
  - Sheller
- Dehydrator
  - Fruit
  - Vegetable
- Drier
  - Grain
  - Hay
  - Hop
  - Nut
  - Seed
- Elevator
  - Grain
  - Roughage
- Ensilage cutter
- Fanning mill
- Feed
  - Carrier
  - Chopper
  - Cooker
  - Grinder
  - Mixer
- Filbert huller
- Flour and cereal mill
- Freight elevator
- Fruit
  - Apple butter maker
  - Cleaner
  - Grader
  - Washer
  - Waxer
- Grain sprouter
- Green feed chopper
- Hay
  - Chopper
  - Crusher
  - Hoist
- Husker—shredder
- Irradiator—seed
- Metal removing magnet
- Molasses heater
- Oat huller
- Pea and bean huller
- Pea viner
- Peach defuzzer
- Potato
  - Grader
  - Seed cutter (Idaho)
  - Washer
- Root
  - Chopper
  - Shredder
- Seed
  - Cleaner
  - Duster
  - Germinator
  - Grader
  - Scarifier
- Storage
  - Fruit
  - Vegetable
- Straw chopper
- Sugar cane press
- Sweet potato
  - Curer
  - Shredder
- Thresher
- Tobacco curer
- Vegetable
  - Grader
  - Tier
  - Washer
- Walnut separator

Dairy

- Agitator
  - Cheese vat
  - Milk can
  - Milk cooler
  - Pasteurizer
- Air conditioner
- Alarm
- Barn cleaner
- Boiler
- Bottle washer
- Bottling machine
- Brine pump
- Bull exerciser
- Butter worker
- Calf weaner
- Cheese maker
- Churn
- Clippers
- Cream
  - Ripener
  - Separator
- Fan
- Feed
  - Carrier
  - Grinder
  - Mixer
- Fence
- Fly
  - Screen
  - Trap
- Groomer
- Heating cable—water pipe
- Homogenizer
- Ice cream
  - Freezer
  - Stirrer
- Insect sprayer
- Irradiator
- Lights
  - Building
- Liquid manure pump
- Litter carrier
- Manure
  - Carrier
  - Drier
  - Pump
- Milk
  - Pump
  - Tester
- Milk cooler
  - Dry
  - Immersion
  - Surface
- Milking machine
- Ox-grub remover
- Pasteurizer
- Prod
- Radiation
  - Infrared
- Radio
- Silage unloader
- Space heater
- Steam accumulator
- Sterile lamp
- Stock tank heater
- Utensil sterilizer
- Ventilator
- Water heater
- Water pail

Farm Shop

- Air compressor
- Alarm
  - Burglar
  - Fire
- Battery charger
- Buffing wheel
- Concrete
  - Mixer
  - Vibrator
- Drill
  - Portable
  - Press
- Emery wheel
- Forge blower
- Glue pot
- Grinder
  - Implement
  - Tool
- Grindstone
- Hoist
- Lathe
  - Metal
  - Wood
- Lights—building
- Paint
  - Brush cleaner
  - Drier
  - Mixer
  - Sprayer
- Planer
- Portable motor
- Pump
  - Gas
  - Grease
- Sander
- Saw
  - Band
  - Buck
  - Circular
  - Cross-cut
  - Hack
  - Jig
- Shaper
- Soldering
  - Iron
  - Pot
- Space heater
- Vulcanizer
- Welder
- Wood
  - Saw
  - Splitter

Field Operations

- Cultivator
- Irrigation
- Manure pump
- Stump
  - Burner
  - Puller
- Tractor
- Truck

Horticulture

- Bleacher—vegetable
- Bulb
  - Drier
  - Grader
- Cider mill
- Cultivator
- Elevator
  - Box or crate
  - Fruit
  - Vegetable
- Fans—orchard
- Fruit
  - Drier
  - Grader
  - Storage
  - Washer
- Greenhouse
  - Fan
  - Heater
  - Ventilator
- Hotbed heater
  - Cable
  - Lamp
- Humidifier
- Insect
  - Sprayer
  - Trap
- Irrigation pump
- Lights
  - Building
  - Greenhouse
  - Plants
- Pollinator
- Seed germinator
- Soil
  - Shredder
  - Sifter
  - Sterilizer
- Soil heating cable
  - Greenhouse
  - Hotbeds
- Soilless culture
  - Pump
  - Solution heater
- Stationary spray plant
- Sweet potato curer
- Vegetable
  - Drier
  - Grader
  - Storage
  - Tier
  - Washer
- Viner—pea
- Water heater

Insect

- Comb wire imbedder
- Fly
  - Screen
  - Trap
- Hive heater
- Honey extractor
- Insect sprayer
- Insect trap
  - Field
  - Fish feeding
  - Home
  - Orchard

Livestock

- Animal prod
- Branding iron
- Brooder
  - Lamb
  - Pig
- Bull exerciser
- Clipper
- Feed
  - Cooker
  - Grinder
  - Mixer
- Groomer
- Hog caller
- Lamb docker
- Lights
  - Building
- Litter carrier
- Manure
  - Carrier
  - Drier
- Meat processing
  - Brine pump
  - Fat cooker
  - Fat cutter
  - Freezer
  - Grinder
  - Hoist
  - Refrigerator
  - Saw
  - Sterile lamp
- Ox-grub remover
- Radiation
  - Infrared
- Sheep shears
- Stock tank heater

Miscellaneous

- Fish barrier
- Limestone pulverizer
- Well drill
- Worm getter

Poultry

- Alarm
  - Burglar
  - Fire
  - Temperature
  - Voltage
- Bone grinder
- Brooder
  - Battery
  - Hover
  - Room
- Caponizer
- Cleaner
  - Dropping board
  - Incubator
  - Poultry house
- Debeaker
- Egg
  - Candler
  - Cleaner
  - Cooler
  - Grader
  - Washer and dryer
- Electrocutioner
- Foot piercer
- Grain sprouter
- Grinder
  - Bone
  - Shell
- Hatcher
- Heating cable—water pipes
- House heater
- Incubator
- Lights
  - Building
  - Poultry
- Manure shredder—grinder
- Mash warmer
- Picker
- Radio
- Scalder
- Singeing device
- Time clock
- Ultraviolet

Self-Excited Electromagnetic Drive for a Resonant Fatigue Machine

ABNER R. WILLSON
MEMBER AIEE

Synopsis: This paper describes the development of low frequency and self-synchronous electric drive equipment for fatigue life testing machines. These test machines simulate the vibrating loads that result from reciprocating engines, variable air loads, and so forth. The self-synchronous feature is important because the inherent high mechanical Q of the system amplifies some 10 to 50 watts of drive power into 2,000 to 5,000 pounds of vibrating load at frequencies from 7 to 40 cycles per second. Mechanically driven types lack sufficient speed stability to maintain resonance, and this results in poor control of the specimen test load. The power requirements for large loads at high speeds are excessive. An experimental 5,000-pound machine with self-excited electromagnetic drive is described. This electric drive is low in both original and maintenance costs. The specimen load is controlled as simply as the volume on a radio receiver. A 100,000-pound machine is nearing completion and is discussed briefly.

DURING the past few years the testing of alloy metals and structures under vibrating loads has assumed an ever increasing importance, especially in the manufacture of structures subject to the vibration of high power reciprocating engines, variable air loads, and imperfectly balanced rotating machinery. Fatigue testing has been practiced by most structural test laboratories to some extent for many years. The common practice has been to apply the vibrating loads by rotating eccentric weights or mechanically driven vibrating cantilever beams. These mechanical systems have inherent limitations which prevent accurate control of the applied force and vibrating frequency. Maintenance costs have been high, and interruptions because of mechanical failures have been too frequent.

The several types of electric drive systems which have been used, all have had characteristics that were not adequate for the type of fatigue testing described in the foregoing paragraph.

1. One type required an alternator with a complex type of frequency control.[1]
2. Another required that the test specimen be a bar of limited dimensions.[2]
3. A third was a separate drive solenoid motor which was connected mechanically to the specimen and excited by a separate oscillator and amplifier. The load limit was considerably below the requirements.[3]
4. A self-resonant type was adapted to the vibration testing of large beams suspended so as to vibrate at their natural frequencies.[4]

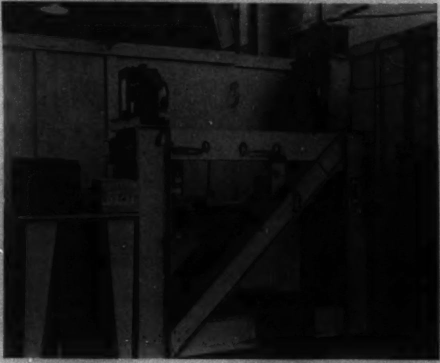

Figure 1. A 5,000-pound fatigue machine, amplifiers, recording meter, actuator solenoid assembly, vibrating beam, and test specimens

In order to design equipment which would meet the requirements of the proposed test program an experimental model was developed which provided a maximum peak-to-peak vibrating load change of 5,000 pounds. The preload may be adjusted to 7,500 pounds which results in a maximum specimen load of 10,000 pounds (refer to Figure 1). A second fatigue machine has been designed and constructed for 100,000-pound maxi-

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Ventilator
Water warmer
Waxer
Wax remelter

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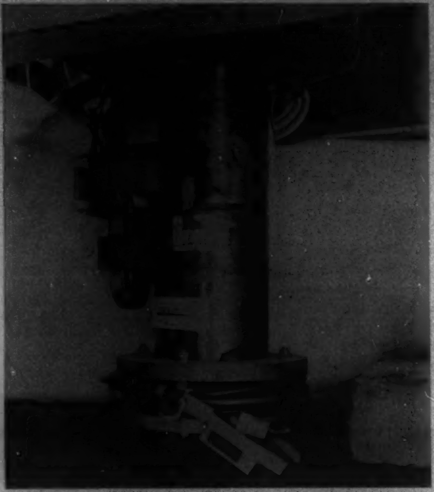

Figure 2. Electromagnetic actuator and pickup, strain gauges and shielded cable, and wedge for indicating vibration amplitude

mum loads. The electric drive is nearing completion (June 1946) but performance data are not yet available. These machines may be adjusted so that the specimen may be vibrated in tension alone, in compression alone, or in a combination of tension and compression.

The Mechanical Problem

The problem was to design and construct a relatively low cost fatigue machine of the resonant type which would provide large vibrating loads and require a minimum of operating supervision and maintenance expense. The resonant type was chosen because of the inherent large amplification of the synchronous drive force into the specimen load.

Rotating beam and other small standard machines were not acceptable because of the limitations in size and shape of test specimens. These machines were designed primarily for test coupons of small dimensions to determine the fatigue life of test specimens of various materials. The requirements for the equipment were not to establish the fatigue lives of present materials but rather to determine the comparative fatigue lives of proposed or actual existing airplane construction members. Typical test problems include the comparative performance of various types of rivets and bolts alone and rivet and bolt patterns on assembled structural sections. Comparative fatigue data also are required for various types of welds and new materials.

The vibrating load demands were to be from a few hundred to 5,000 pounds. These loads would be the peak-to-peak loads as shown in Figure 11. That is, the specimen would be preloaded statically in excess of half the total vibrating load, and the resulting maximum and minimum specimen loads then would be the preload plus or minus one-half of the peak-to-peak vibrating load change. For the example shown in Figure 11 the preload was 2,350 pounds, and the vibrating load change was 3,766 pounds. The maximum specimen load was 4,233 pounds, and the minimum was 467 pounds.

The 5,000-Pound Machine

The development model of the Boeing fatigue machine is illustrated in Figure 1. The vibrating beam is a pair of 6-inch commercial steel channels supported at the right end by flexure pivots. The beam is vibrated by the actuator solenoid assembly shown at the left end of the beam in Figure 1 and also in Figure 2. This actuator is in two sections. The top half is bolted to the framework, and the lower part is fastened to the vibrating beam with four one-quarter-inch aluminum-alloy bolts on which are mounted strain gauges.[8] Copper tubing electrostatically shields and mechanically protects the gauges and their leads. The shielded cable shown coiled around the lower shell connects the gauges to a

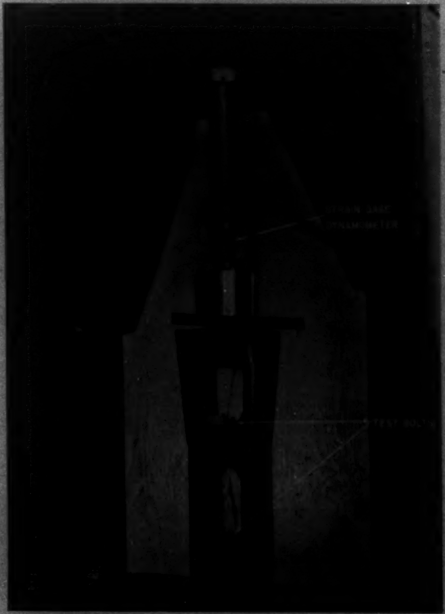

Figure 3. Test bolts and strain gauge dynamometer

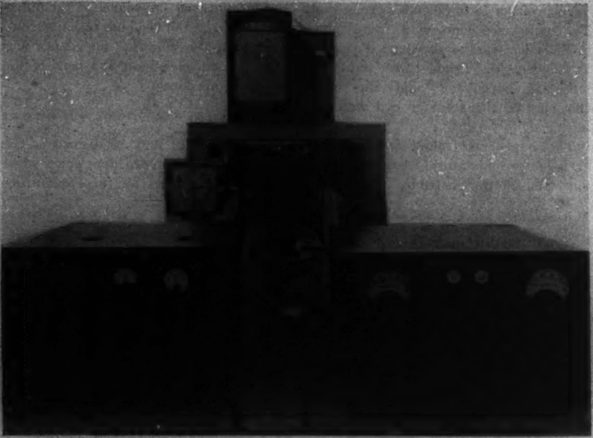

Figure 4. Amplifiers for 5,000-pound fatigue machine

The recording amplitude meter shows the preliminary adjustment of preload and adjustment of impedance matching transformers. The machine was operating with a specimen load of about 3,000 pounds

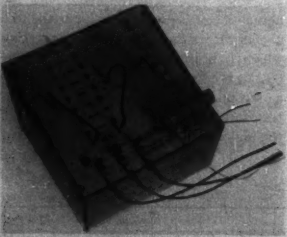

Figure 5. Impedance matching transformer for 5,000-pound fatigue machine

vacuum tube voltmeter and recording oscillograph to indicate and record the driving force.

The test specimen is attached between the fixed support shown at the upper right in Figure 1 and a point on the vibrating beam near the flexure joint to give a 9.5-to-1 mechanical advantage. The test specimens illustrated in Figures 1 and 3 are 5/16-inch special alloy steel bolts. The bolt heads and nuts are shown at the bottom and top of the O jigs. Strain gauges are mounted on the top adjustment bar. These gauges are for measuring the loads in the specimen, both static and dynamic. These specimens were preloaded with the flat spring and turnbuckle shown in the lower part of Figure 1. The preload is measured statically by the strain gauge output into a d-c bridge-type instrument which is calibrated directly in pounds per square inch. This instrument with its white dial is shown at the lower center in Figure 12. The recording strip chart meter shown at the top of Figure 4 indicates the amplitude of beam vibration and is therefore an indication of the load in the specimen. The broken curve on the strip chart shows the preliminary calibration of the meter and adjustments of the amplifier output and impedance matching for the bolt test. The exciting voltage is generated by the pickup (Figure 2), amplified by the equipment shown in Figure 4, and fed through the impedance matching transformer (Figure 5) into the actuating solenoid (Figure 2). Each of these electric components is to be described.

The Electric Circuit

The electric circuit may be considered as that of an oscillator with the output sufficiently amplified to drive the vibrating beam (refer to Figures 2, 4, and 6). The exciting voltage is generated by a permanent magnet attached to the vibrating beam and vibrating into fixed coils (Figure 2). The pickup voltage is usually in the order of three to six volts. It is fed to the input circuit of the amplifiers (Figure 4). The amplifier output is impedance matched to the actuator winding by a special transformer designed for the 7- to 40-cycle-per-second frequencies (Figure 5). The polarizing coils are energized by the rectified power supply for the 6L6 amplifier. A filter separates the polarizing current from the plate voltage supply to the amplifier tubes.

The recording meter, Figure 4, indicates the amplitude of the vibration of the beam. It is a 0–1.4-volt 1,400-ohm d-c voltmeter. It is connected to the output of the pickup coil through a series calibrating resistor and a small dry rectifier. The meter reading follows the amplitude of the beam vibration with sufficient range and accuracy for any one test frequency and load condition. The strip chart gives a continuous record of specimen load and operating time. The vibrating specimen loads first are measured with strain gauges and an oscillograph. The recording meter then is calibrated for that test condition. The record on the strip chart meter shows a gradual decrease slightly before the specimen fails.

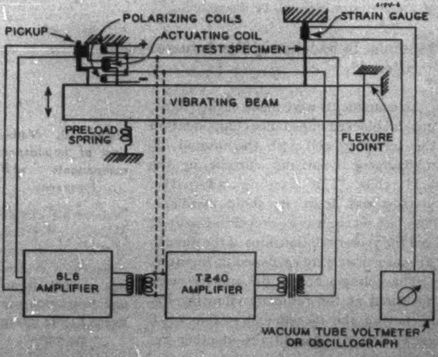

Figure 6. Electric circuit of the 5,000-pound fatigue machine

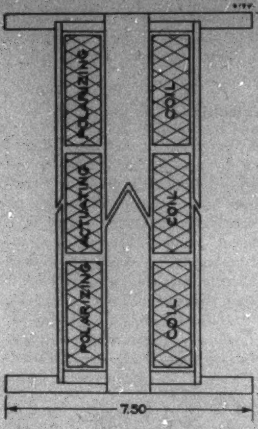

Figure 7. Cross section of the actuator solenoid assembly

After the failure of the specimen the beam falls on a limit switch which opens all electric circuits including the electric drive on the strip chart.

The Actuator System

A cross section of the actuator system is shown in Figure 7. A polarizing magnetic field is induced by the top and bottom coils. They are wound with number-30 enameled magnet wire and usually are worked at about 2,500 ampere turns. The central coil corresponding to the voice coil in a loud speaker, supplies the actuating force. This coil, the load for the output of the amplifier, usually is worked at an alternating current density of 150 to 350 milliamperes and at frequencies between 12 and 30 cycles per second. The impedance characteristics of the actuator coil will be discussed.

Motional Impedance of the Actuator Solenoid

Measurements were made of the vibrating and blocked impedance components of the actuating coil with the normal 50-milliampere polarizing current in the field coils. The exciting alternating voltage was from an audio oscillator. The test set up was as shown in Figure 8, and the vector relationships of the quantities measured were as shown in Figure 9.

The voltages across *BD* and *ED* were measured at the resonant vibrating frequency of the mechanical system (19.6 cycles per second) and also at other frequencies from 18 to 50 cycles per second. The impedance components were calculated and plotted as shown on the curves in Figure 10. From these data the motional impedance[8] and the drive efficiency of the actuator coil were determined.

The impedance values for the 19.6-cycle-per-second natural mechanical resonant frequency are listed in Table I.

Motional impedance = 365 ohms

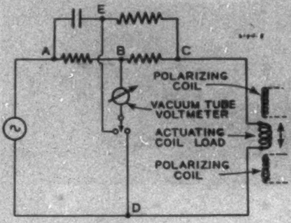

Figure 8. Setup for measuring motional impedance

For vector relations refer to Figure 9

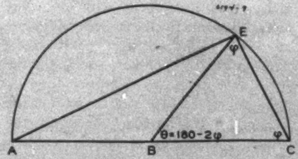

Figure 9. Vector relations of the voltages in Figure 8

The motional impedance is the difference between the 666-ohm resistive component of the vibrating impedance and the 301-ohm resistive component of the blocked impedance.

Efficiency (666 − 301)/(666) = 55 per cent

This efficiency may be defined as the ratio of output power to input power when the load resistances for these quantities are considered as

$$\frac{\text{Vibrating resistive component} - \text{blocked resistive component}}{\text{vibrating resistive component}} = 55 \text{ per cent}$$

The sharp resonance curves show the importance of operating the electric drive at the mechanical resonant frequency. The power into the actuator coil is obviously a maximum when the motional resistive component is a maximum.

Amplifier

The amplifier is in two sections. The first has a 6J5 tube input, and a push-pull 6L6 output with transformer coupling between the 6J5 plate and the 6L6 grids. When the specimen loads are less than 2,000 pounds the output of the 6L6 plates is fed through a special impedance matching transformer directly to the actuator coil. For greater loads the 6L6 output drives the second amplifier section which has push-pull TZ40 tubes. The output is coupled to the actuator through a second special impedance matching transformer (Figure 5). The output of the 6L6 amplifier directly into the load at 12 to 30 cycles per second is 15 to 30

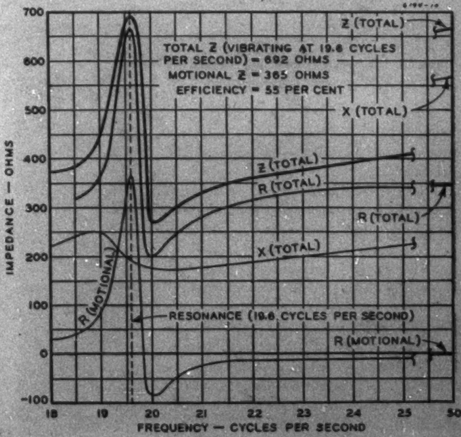

Figure 10. Variation of impedance components with frequency

Because the driving power is a function of the resistive component it is evident that a small departure from resonance means a large loss in power watts. For greater specimen loads the TZ40 amplifier delivers 30 to 90 watts at the same frequencies. In this case the 6L6 amplifier supplies only the three to six watts required to drive the TZ40 grids.

Low Frequency Transformers

When one considers the comparative sizes of conventional power transformers of equal kilovolt-ampere and voltage ratings to operate on 400-, 60-, and 25-cycle circuits it will be evident that the 7- to 40-cycle power requirements of the vibrating machine demand transformers with much copper and iron in order to face both the tube output and the actuator input circuits with sufficient inductance to transfer the low frequency energy with reasonable transformer losses and proper impedance matching.

The impedance matching transformer illustrated in Figure 5 is used to match the TZ40 push-pull plates to the actuator coil. For vibrating the specimens shown in Figure 3 the plate-to-coil impedance ratio was 59 to 1. The frequency for this test was 12.5 cycles per second.

Figure 12. A 100,000-pound fatigue machine

The vibrating beam weighs 1,100 pounds and the actuator coil assembly weighs 440 pounds; five-eighths-inch heat treated bolt specimens are being fatigue tested

Performance

A typical test setup is shown in the illustrations. The test specimens were 5/16-inch special alloy bolts. The initial preload and vibrating load in the specimen was measured by the output of the specimen strain gauges. The force produced by the actuator solenoid was measured by a second set of strain gauges. The oscillograph curves reproduced in Figure 11 show the static preload and the magnitude and phase relation of the force supplied by the electric drive system and the resulting vibrating load change in the test bolts. The over-all Q of the system was 100. This 100-to-1 ratio of the 3,766-pound load in the specimen bolts to the 27.7-pound driving force results from the 9.5-to-1 mechanical advantage of the application points of the driving force and specimen load and the amplification which results from operating the beam system at its natural resonant frequency.

Table I. Impedance of the Actuator Solenoid

Component	Impedance, Ohms
Total vibrating, 15.7° phase angle	692
Resistive	666
Reactive	184
Total blocked, 34.5° phase angle	365
Resistive	301
Reactive	206

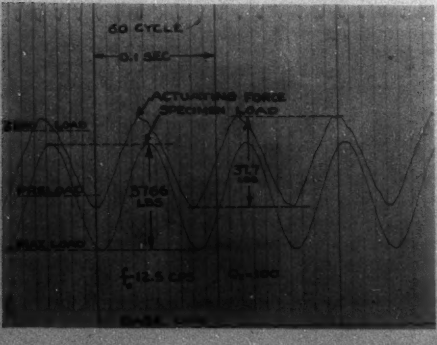

Figure 11. Oscillograph record of the actuating force and resulting load in the specimen

A Q of 100 was obtained for this test at 12.5 cycles per second

The lead weight shown clamped to the beam in Figure 1 lowered the natural frequency of the beam system as it was moved away from the pivot point. Lowering the frequency increased the specimen load but decreased the transformer efficiency. The optimum frequency for this test condition was 12.5 cycles per second.

Measurements were made of the various voltages and currents while operating with a 3,000-pound load change in the specimens. These rms voltages and currents are listed in Table II. When operating with a 4,000-pound load in the specimens the actuator alternating current was 250 milliamperes.

The 100,000-Pound Vibrating Machine

A vibrating machine was designed for a maximum specimen load of 100,000 pounds. This is 50,000 pounds vibrating load change and 50,000 pounds preload. The mechanical portion has been completed and temporarily is being driven for light specimen loads by a mechanical eccentric until completion of the electric drive (Figure 12). Three five-eighths-inch bolts were being tested in the O jigs at the left. Directly above the specimens is shown a copper cylinder which electrostatically shields and also mechanically protects the strain gauges which measure specimen loads. The static prestress is measured on the d-c

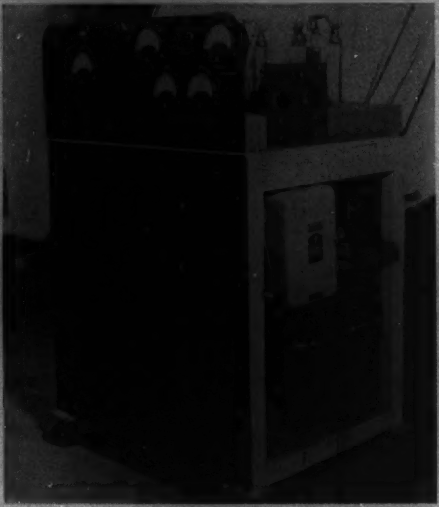

Figure 13. Amplifiers for 100,000-pound fatigue machine

Table II. Voltages and Currents at 3,000-Pound Load Change

	Rating
Pickup voltage	5
Coupling transformer primary voltage	150
Coupling transformer secondary voltage	78
Output transformer primary voltage	530
Output transformer secondary (load) voltage	84
Load current (ma)	130
Polarizing current (ma)	40

bridge shown at the central rear. Dynamic loads are measured with a vacuum tube voltmeter and oscillograph as in the measurements for the 5,000-pound machine. The lower specimen mounting is fastened near the pivot point of the 1,100-pound offset beam. The 12-inch cylindrical sections which house the polarizing and actuator coils are shown at the right end of the beam. The top section vibrates with the beam, and the lower section is fastened to the framework. A large helical preload spring is mounted directly above the actuator assembly and is adjusted by the hand crank shown at the upper right. The preload also may be adjusted by the specimen adjusting bolts shown at the upper left.

The actuator assembly consists of the 12-inch steel shells, the enclosed polarizing coils at the top and bottom, and the actuating coil at the center. The weight of the copper wire is 170 pounds, and the steel shell and core assembly weighs 270 pounds. The polarizing coils will be operated at 2,000 to 5,000 ampere turns, and for maximum load conditions the actuator is expected to require about 400 watts. A permanent magnet and a solenoid attached respectively to the stationary and moving shell sections will furnish the synchronous exciting voltage to the amplifier input. The driver amplifier is a 6L6 type similar to the one described for the 5,000-pound machine. The power amplifier stage contains push-pull RK63 power tubes rated at 200 watts plate dissipation. Special low frequency transformers couple the two amplifiers and match the output of the power stage to the actuator solenoid. Preliminary tests have indicated that the operating frequencies will be from 7 to 40 cycles per second.

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Application of Compandors to Telephone Circuits

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Synopsis: Prior to the war a compandor was developed for general use on telephone circuits of the Bell System. This equipment provides power range compression at the transmitting end and power range expansion at the receiving end of a telephone channel. The resulting crosstalk and noise reductions of about 25 decibels are large enough to make useful circuits out of channels which otherwise would be unworkable. This paper considers engineering problems in the application of compandors and describes the circuit principles involved in the equipment.

GOOD transmission characteristics of a communication channel are primarily the result of care exercised in the original design and construction of the communication system of which the channel is a part, whether it be cable, open wire, radio, or mixtures of these. In the designs allowance is made for growth, so that future added channels will meet noise and crosstalk limits themselves and at the same time not react on the earlier installations. There are economic limits to such anticipatory design, however, and sometimes demands for additional channels arise more rapidly than was expected. It then may prove either physically difficult or expensive to provide the relief needed. For example, the obvious measure of adding a 12-channel carrier system to another pair of wires of an open-wire line might require elaborate reassignment of all the pairs in the lead, retransposition, and retensioning to avoid undue crosstalk from system to system.

The compandor to the present time has found its main field of use in meeting such situations. The principles involved are sufficiently general, however, to suggest that circuit designers will find it advantageous to consider the application of compandors, in some form, in the over-all planning of new systems. It then becomes a question of economics as to whether this, or other methods, or both should be employed to meet noise and crosstalk limits.

The compandor described is a voice-frequency device, applied channel by channel where needed, which enables noise and crosstalk reductions up to about 25 decibels to be obtained. Its first application to wire lines in the Bell System was made early in 1941 on the Charlotte–Miami and Charlotte–West Palm Beach routes. It proved of value during the war in permitting rapid additions to cope with some of the demands stimulated by military installations. Circuits which otherwise would have been unworkable were brought to satisfactory standards. The number of compandors in use at the present time is over a hundred, on the routes mentioned and in other locations, and substantial additions are to be made during the current year.

A compandor was applied to the long-wave New York–London radio circuit some years ago.[1,2] At that time vacuum tube circuits were utilized for compression and expansion, but the general theory stated then still is applicable to the present compandors, in which there are vario-lossers formed of copper-oxide varistors. An experimental model using varistors was installed in 1938 on the Boston–Provincetown circuit, which is partly wire and partly radio; this is still in operation. Other applications of compandors outside the Bell System have been made by the British Post Office.[3]

Functional Description

Circuit Application

A compandor consists of two units, a compressor and an expandor, from which it obtains its name. Each is a piece of equipment of about one-half cubic foot in volume. The units are one-way devices operated at voice frequencies so that they are installed, as Figure 1 indicates, on the 4-wire side of the circuit. The 4-wire circuit is converted to 2-wire by the standard hybrid coil and balancing network arrangements indicated. The compressor is installed at the transmitting end of a one-way channel and the expandor at the receiving end.

The line between the two terminals may be voice frequency or carrier and it may include a radio link. The channels may be equipped with compandors in each direction of transmission, or only in one direction, as required. The functions of the attenuating pads P_C and P_E and of the cross controls shown in Figure 1 will be discussed later.

Compression and Expansion

The compressor and the expandor complement each other; what one does the other reverses. If the compressor inserts amplification in the channel the expandor inserts an equal attenuation. The noise reduction afforded by these complementary units comes about because the compandor allows a basic principle to be pushed further than is possible otherwise. This is the old proposition that if the transmitting level of a circuit is increased, a loss can be inserted at the receiving end without changing the net loss, but reducing noise coming into the line between terminals by the amount of the inserted loss. Usually transmitting levels cannot be increased without overloading the line amplifiers by the stronger signals. The compandor avoids this difficulty by giving no gain, or even a loss to the strongest signals, and by giving the maximum gain to the weakest signals, which although so amplified are well below the overload point.

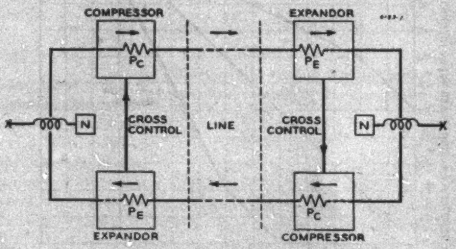

Figure 1. Compandors applied to 4-wire telephone circuit

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The variable gains and losses in the compressor and expandor are controlled automatically by the syllabic envelope of the signal itself, which ordinarily is speech. There are rapid variations in speech power and these produce an incessant variation in the gains and losses of the two units. It is simpler to consider the action of the compandor in terms of steady-state testing power.

Figure 2 shows input-output characteristics of the compressor and expandor, measured separately. The input power was applied in each case at the toll switchboard or other zero-level points ahead of the device. The output power was measured at a zero-level point beyond the device. (Zero-level points in a toll circuit are those points at which the net loss as measured from the toll switchboard is zero; these points are designated before insertion of the compandor.)

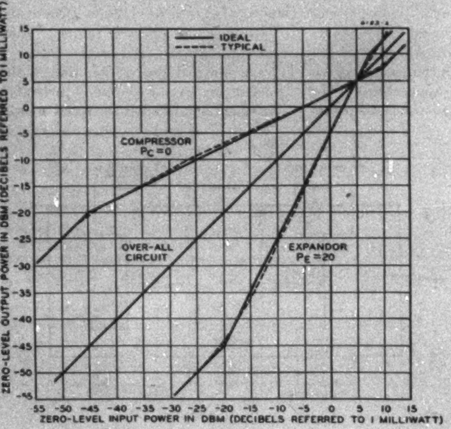

Figure 2. Input-output characteristics of compressor and expandor

The 45-degree slopes at the extremes of the curves mean that the devices are ordinary amplifiers at those values. The companding action takes place in the range between called the companding range. In this range the slope of the compressor curve is one-half. A range of input power of 55 decibels (minus 45 to plus 10 "dbm" is converted by the compressor to an output range of 27.5 decibels. (This electric power is expressed in decibels referred to the commonly used base of one milliwatt. In the Bell System this unit generally is abbreviated "dbm" and for the sake of convenience the abbreviation will be used throughout this paper.) The slope of the expandor curve is two, and it converts an input range of 27.5 decibels to an output range of 55 decibels.

POWER LEVELS

In the usual communication circuit two input powers, differing by 30 decibels, for example, will remain 30 decibels apart from one end of the circuit to the other. This condition is changed when a compandor is inserted, as is shown in Figure 3. This indicates graphically how a range of input power is reduced by the compressor for transmission and then restored by the expandor.

In Figure 3 the power levels and level changes in a compandored circuit are shown at three points.

1. A zero-level input point.
2. A zero-level output point.
3. A point between the compressor and expandor which would be a zero-level point if the devices were not there.

One input power only (for this compandor it is plus five "dbm") goes through the system unchanged. The top of the companding range, plus ten "dbm" is attenuated by 2.5 decibels in the compressor and thus is transmitted over the line at a level lower than normal. Input powers between plus five and minus 45 "dbm" are amplified by amounts equal to half the separation in decibels between the input power and plus five "dbm." Zero-"dbm" testing power, for example, is amplified by 2.5 decibels and a power of minus 30 "dbm" by 17.5 decibels. Powers that were 30 decibels apart at the input thus are transmitted 15 decibels apart. Powers at the bottom of the compounding range, minus 45 "dbm," and below are amplified 25 decibels. The whole companding range, 55 decibels, thus is transmitted as a range of 27.5 decibels. At the receiving end the expandor restores the original separation between the powers.

Since the maximum gain of the compressor is 25 decibels, the maximum loss of the expandor is also 25 decibels, and this is the maximum possible noise reduction. Ordinarily the crosstalk reduction is slightly less than this figure, but under some circumstances to be discussed later it may be considerably more. More noise reduction could have been obtained if the value of power which receives neither gain nor loss had been put at the top of the companding range, for then the maximum loss would have been 27.5 decibels, one-half the whole range. This was found to be undesirable for it increased the frequency of occurrence of high speech peaks on the line, thus reducing the overload margin.

Noise and Crosstalk Reduction

When a compandor is installed on a circuit to reduce noise, crosstalk, or both, adjustments can be made to suit various circumstances by selecting the values of the attenuation pads P_C and P_E indicated in Figure 1. In the compandor described in this paper, the sum of these losses ($P_C + P_E$) must be kept at 20 decibels to maintain the complementary relation between compressor and expandor. Thus there is only one independent adjustment to be made.

Formulas describing the relations between steady-state input and output powers are helpful in understanding the basis for selecting the values of P_C and P_E. Referring to Figure 2, the slope for the straight line representing the compressor characteristic is one-half, and the output power for an input of 0 "dbm" is 2.5 "dbm." The pad P_C of the compressor will reduce further the output power. Thus, in "dbm," the output power C_{OUT} is related to the input power C_{IN} as follows:

$$C_{OUT}=\frac{1}{2}C_{IN}+2.5-P_C \qquad (1)$$

The slope for the expandor is two, and the output power for an input of zero "dbm" is minus five "dbm" when the pad P_E is 20 decibels. If the value of P_E is decreased from 20, the output will be increased by twice the change in P_E, since P_E is in the input side. Thus the output power E_{OUT} is related to the input power E_{IN}.

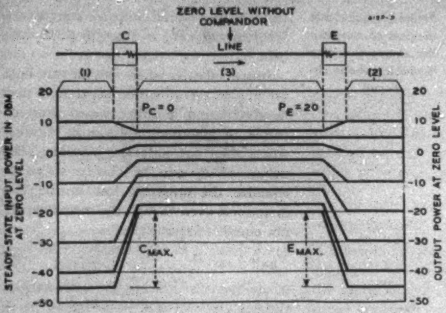

Figure 3. Power levels in telephone circuit with compandor

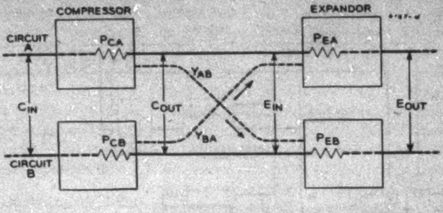

Figure 4. Losses in crosstalk paths between telephone circuits with compandors

$$E_{OUT}=2E_{IN}-5+2(20-P_E) \qquad (2)$$

As stated in the foregoing text, to preserve complementary action,

$$P_C+P_E=20 \qquad (3)$$

Thus, choosing P_C for the independent adjustment, the expression for E_{OUT} becomes

$$E_{OUT}=2E_{IN}-5+2P_C \qquad (4)$$

If E_{IN} is made equal to C_{OUT}, by substitution from equation 1 it will be found that $E_{OUT} = C_{IN}$; the over-all net loss is unaffected.

The maximum gain inserted by the compressor occurs when $C_{IN} = -45$ "dbm." At this value $C_{OUT} = -20 - P_C$, and the gain $C_{OUT}-C_{IN}$ at this value is $(25-P_C)$ decibels. This maximum gain is matched by the complementary loss in the expandor, so that the maximum loss in the expandor, designated E_{MAX}, is

$$E_{MAX}=25-P_C \qquad (5)$$

Noise Reduction

If noise is induced in the line between its terminals, the maximum loss is desired at the receiving terminal. To obtain this, the compressor pad P_C is made zero, and the expandor pad P_E is made 20 decibels. The power levels then correspond to Figure 3. In general, the noise advantage is E_{MAX}; thus, if for any reason P_C is not zero, the noise advantage becomes $(25-P_C)$ instead of 25 decibels.

There is an upper limit to the amount of noise with which the expandor can deal. If the noise is high enough it will operate the expandor controls and the loss inserted will be less than 25 decibels. The upper limit is quite high, however. As Figure 2 shows, any noise corresponding to a power of minus 20 "dbm" or less will not operate the expandor controls and therefore will be attenuated by the full loss. A single frequency noise of minus 20 "dbm" would read plus 70 decibels on a Western Electric type 2 noise measuring set,[4] unweighted. The corresponding figure for noise spread over the frequency band, such as static, is about 55 to 60 decibels above reference noise, with message weighting. Noise of this value also is reduced by the expandor to 30 to 35 decibels at the zero level, or 20 to 25 decibels at a minus 10-decibel level, which is within satisfactory limits.

Crosstalk Reduction

The crosstalk problem usually differs from the noise problem in that whole groups of channels need to be considered in relation to each other. These may be, for example, all of the channels of the same frequency allocation and direction of transmission in several carrier systems in the same cable. Nothing is gained in this case by raising all of the transmitting levels and inserting compensating losses at the receiving end, since the couplings between channels are not changed.

The crosstalk advantage provided by the compandor is a result of its selectivity with respect to amplitude. The range of talking volumes applied to a telephone circuit is about 40 decibels; of these, it is the louder ones which constitute the sources of crosstalk. The compressor can be arranged to discriminate against these by lowering the level for high talking volumes while raising the level for low talking volumes. The compressor therefore can be used to control crosstalk into other channels. A further crosstalk reduction is provided by the expandor on the receiving end of the disturbed circuit. This can be used to control crosstalk from other channels. Ordinarily it is desirable to gain the maximum advantage at the expandor end, since this also provides the maximum noise reduction. In special cases it may be more useful to work the compandor in such a way as to gain a large advantage at the compressor end. For example, the total advantage may be divided between two circuits. The choice of values for the compressor and expandor pads P_C and P_E depends on such considerations.

Table I

No. of Channels Being Equipped	Value of PC For Which CXA = 0
1	5
5	4
40	2
All	0

Compressor Crosstalk Advantage

The compressor crosstalk advantage CXA for test tone is the reduction in the power transmitted to the line caused by the compressor, or

$$CXA \text{ (for test tone)} = C_{IN}-C_{OUT} = \frac{1}{2}C_{IN}-2.5+P_C \qquad (6)$$

For input powers less than plus five "dbm" this is negative, that is, a disadvantage unless P_C is made greater than zero.

In speech the power varies widely during the time one talker is speaking, and the averages for different talkers cover a wide range of volumes. Also there are often several possible disturbing sources. For these reasons the compressor crosstalk advantage for speech requires a statistical evaluation. This has been done, with the results listed in Table I.

The number of channels being equipped signifies the quantity which have such frequency allocations as to be potential crosstalk disturbers, and in which compandors are being installed.

Designating the figures in the right hand column as P'_C, the crosstalk advantage formulation for speech then becomes

$$CXA \text{ (for speech)} = P_C-P'_C \qquad (7)$$

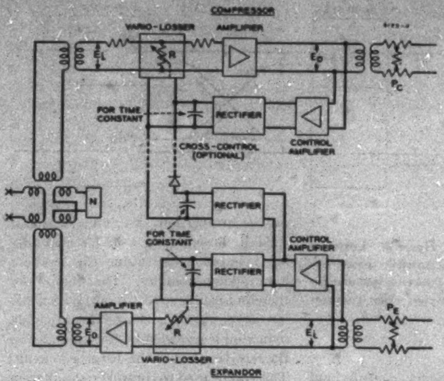
COMPRESSOR
VARIO-LOSSER
AMPLIFIER
R
E_L
E_O
P_C
FOR TIME CONSTANT
RECTIFIER
CROSS-CONTROL (OPTIONAL)
CONTROL AMPLIFIER
N
RECTIFIER
FOR TIME CONSTANT
CONTROL AMPLIFIER
RECTIFIER
AMPLIFIER
P_E
E_O
R
E_L
VARIO-LOSSER
EXPANDOR

Figure 5. Schematic diagram of compandor at one end of a 4-wire telephone circuit

In ordinary practice, P_C is made equal to P'_C so that CXA is zero, that is, the circuits with compandors are no more disturbing to other circuits than previously. In special cases, it may be desirable to reduce the disturbing effect to other circuits and this is done by making P_C greater than P'_C.

Expandor Crosstalk Advantage

Since crosstalk of manageable proportions is too weak to operate the expandor controls, the expandor crosstalk advantage is the maximum loss in the expandor E_{MAX}, or

$$EXA = 25 - P_C \tag{8}$$

The formula shows that it is desirable to make P_C as small as possible. As previously indicated, however, this is limited by the disturbing effect of the circuit to which the compandor is applied.

Total Crosstalk Advantage

Total crosstalk advantage from one circuit to another is the sum of the compressor crosstalk advantage CXA of the disturbing circuit and the expandor crosstalk advantage EXA of the disturbed circuit. In the usual cases with P_C made equal to P'_C, the value of CXA is zero, so that the total crosstalk advantage from circuits either with or without compandors to a circuit with a compandor is EXA, or $25 - P'_C$ or from 20 to 23 decibels.

In special cases where specific couplings are known the crosstalk advantages may be distributed in other ways. For example, assume a situation as in Figure 4 where there are only two circuits. The values of P_C can be chosen so as to equalize the crosstalk loss in the two directions, even though the couplings are not equal. Assume that

$$Y_{BA} = Y_{AB} + D \tag{9}$$

where the Y's indicate the attenuations through the crosstalk paths from B to A and A to B, and D is the difference between them. In the A–B direction the total advantage is

$$CXA_A + EXA_B \tag{10}$$

and in the B–A direction the subscripts are interchanged. It is desired to have

$$CXA_A + EXA_B = CXA_B + EXA_A + D \tag{11}$$

so that a greater advantage in the A–B direction will compensate for the fact that there is less attenuation to crosstalk in that direction. Using equations 7 and 8, with added subscripts to denote terminals,

$$P_{CA} - P'_C + 25 - P_{CB} = P_{CB} - P'_C + 25 - P_{CA} + D \tag{12}$$

Therefore

$$P_{CA} - P_{CB} = \frac{D}{2} \tag{13}$$

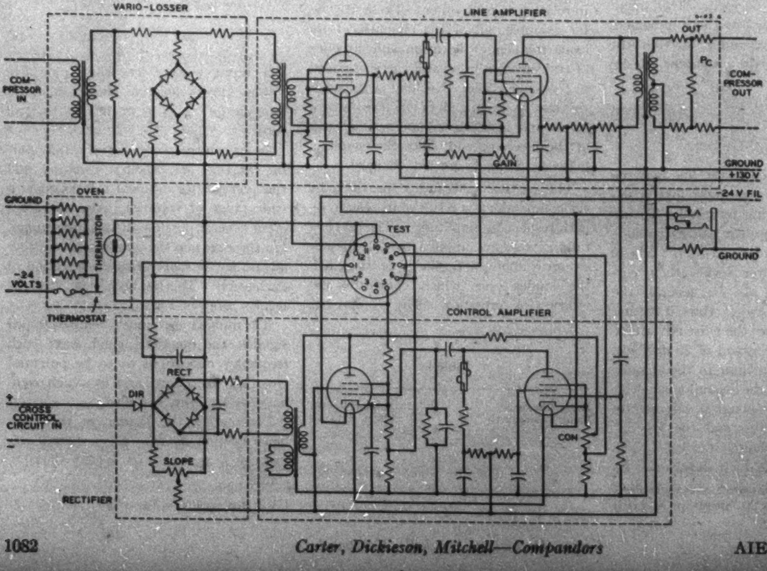
VARIO-LOSSER
LINE AMPLIFIER
OUT
P_C
COM-PRESSOR IN
COM-PRESSOR OUT
GAIN
GROUND
+130 V
-24 V FIL
GROUND
OVEN
GROUND
THERMISTOR
-24 VOLTS
THERMOSTAT
TEST
CONTROL AMPLIFIER
RECT
DIR
CROSS CONTROL CIRCUIT IN
SLOPE
RECTIFIER
CON

Figure 6. Schematic diagram of compressor, Western Electric type 1A

Assume, for example, that $D = 20$ decibels. Since there are only two circuits and both are equipped, $P'_C = 0$, so that P_{CB} can be made zero. Then, by equation 13, P_{CA} should be ten decibels. The corresponding values of P_{EB} and P_{EA} are 20 and 10 decibels, respectively. The total advantage in the A–B direction is $10 + 25 = 35$ decibels, while that in the B–A direction is $0+15=15$ decibels. The same formula (13) also can be derived from equations 6 and 8 for test tone. It is of some interest that in a situation of this kind the total advantage in one direction can exceed E_{MAX}, since the advantage is the sum of the value of CXA on one circuit and EXA on the other.

Transmission Considerations

Net Loss Variations

Net loss variations in a communications channel are deviations from the nominal or assigned value of over-all net loss. These may occur from a variety of causes, but on wire lines primarily from fluctuations of weather. A positive variation, which increases the net loss, is undesirable because the loudness of the speech heard by the subscribers is reduced. Negative variations, though improving the loudness of the speech, also may be undesirable if too great. Speech reflected from the far end of a circuit then may be audible as an annoying echo. If the negative variation is extreme the circuit finally may become unstable and sing.

The compandor accentuates the problem of net loss variations. Equation 4 shows, for powers within the companding range, that if the input power to the expandor is changed by a line loss variation, the change in output power is twice as great. For this reason, compandors are applied to most advantage on circuits which have good regulation of net loss.

Effect of Cross Control on Echo and Singing

When both sides of a 4-wire circuit are equipped with compandors the effect of the increased net loss variations on the round trip path can be avoided by means of cross controls from the expandor to the compressor at each terminal. These are indicated in Figure 1. By means of the cross control, when the expandor is inserting gain in its own path it simultaneously inserts an equal loss in the compressor at the same terminal. This is a unilateral action, since a directional element prevents the control current of the compressor from reaching the expandor. Bilateral controls also have been tested, but were found to be unnecessary. However, in most applications to modern well-regulated carrier circuits the unilateral control is not needed either, so that its use is made optional.

When cross controls are used in a 4-wire circuit equipped with compandors in both directions, the echo conditions are better in some circumstances than on the same circuit without compandors. This advantage is most pronounced when the compandors are at the terminals and the circuit is not switched to other circuits to distant cities at either end. The echo then returns so quickly that it passes through the compressor while the cross control from the expandor at the same terminal is inserting a loss in it, and the net loss to the echo in the return path thereby is increased as compared with a return path not having a compandor. If there is a long connecting circuit the echo may return after such an interval that the cross control has ceased its action, so that this echo improvement no longer is obtained.

With respect to singing, the cross control tends to cancel the effect of the increased net-loss variations introduced by the expandor, since a compensating loss or gain is inserted in the return path.

Quality Considerations

The Compandor, a Dynamic Device

For simplicity, the preceding sections have described the properties of the compandor in terms of steady-state testing power. Its primary purpose, however, is to transmit speech, which is a constantly varying quantity. The wide fluctuations of power during speech cause continual variation of loss in both compressor and expandor. In these circumstances it was not easily predictable that speech powers could be compressed, transmitted over a telephone line of limited frequency range,

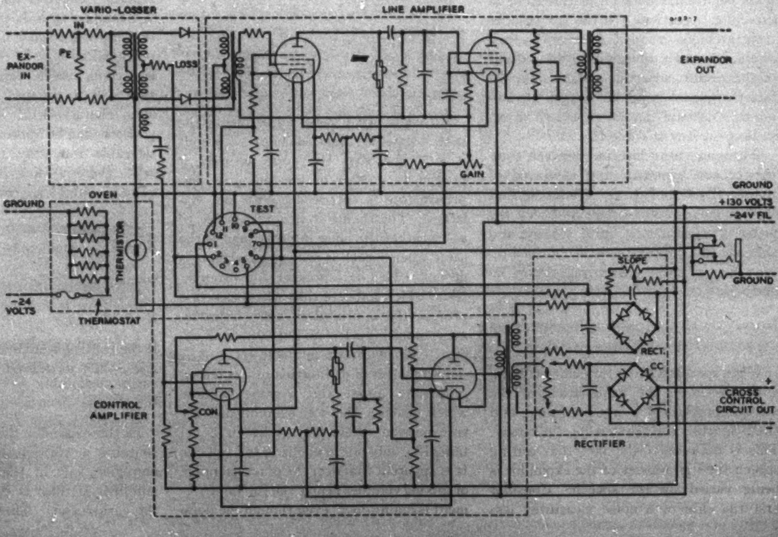

Figure 7. Schematic diagram of expandor, Western Electric type 1A

and then themselves guide the expandor in such a way as to reproduce the original dynamic pattern sufficiently well to be acceptable to the ear.

The quality characteristics depend on close matching of the characteristics of the vario-lossers at the two terminals, and on the selection of the proper attack and recovery times in the control circuits. Attack time is the interval required, after the sudden application of a steady testing signal, for the gain of the compressor or of the expandor to reach substantially its final value. Recovery time is the analogous interval after the interruption of a steady testing signal. If the attack and recovery action is too abrupt, modulation products may be created which may be heard as distortion and noise. If the attack time is too slow there may be mutilation of the initial part of a syllable. If the recovery time is too slow the loss of the expandor is not inserted fully between syllables. Talking and listening tests established that attack times on the order of ten milliseconds and recovery times on the order of 50 milliseconds gave satisfactory quality on good toll circuits.

Noise Advantage in the Absence of Speech—Aural Tests

When speech is not being transmitted, the expected improvement, as discussed in preceding sections, is equal to the loss inserted by the expandor. Measurements of this improvement have been made with a noise measuring set and also by ear balance comparisons for noises both below and above the threshold of expandor operation. In the ear balance tests, each of a group of observers listened alternately with the expandor in and out of the circuit, adjusting the noise in the latter condition until he judged the two noises to sound equally loud. The results are shown in Table II.

The noises were inserted between compressor and expandor and measured at zero-level points. It is of interest that the ear balance measurements check the meter measurements quite closely and that the full improvement is obtained up to values of noise measuring nearly 60 decibels.

Noise Advantage in the Presence of Speech—Aural Tests

When speech is being transmitted along with the noise the explanation of the noise improvement is not so simple, and the figures are somewhat different. This is because of the fact that during speech the vario-losser of the expandor is being varied by the speech. Consider first the case of a noise measuring less than 60 decibels above reference noise at a zero-level point. Between syllables the expandor attenuates this 25 decibels. During a syllable the speech operates the expandor, removing loss, perhaps to the extent of 25 decibels. Thus the attenuation given the noise during speech varies according to the strength of the syllables. If the recovery times of the control circuits are not chosen carefully the variation of the background noise may be quite audible; from its nature this effect has been named "hush-hush." The recovery times chosen disguise the hush-hush on a message circuit from all but the most critical listener.

Table II. Noise Improvement—Compandor In Versus Out

Magnitude of Noise, Db			Noise Improvement, Db	
	With Compandor			
Without Compandor	Noise Measuring Set	Ear	Noise Measuring Set	Ear Balance
49	24		25	
58	33	32	25	26
68	47	48	21	20
77	65	66	12	11

Table III. Noise Improvement in the Presence of Speech

	Ear Balance Tests		
	Zero Level Noise, Db		
Talking Volume at Zero Level, vu	Without Compandor	With Compandor	Noise Improvement, Db
No speech	58	32	26
−10	58	34	24
+ 8	58	37	21

Although with these adjustments the speech syllables mask the noise variations it is a fact that the noise actually is present in greater or less amounts and presumably is interfering with the syllables. That being so it is reasonable to ask whether the noise improvement is the same when speech is present as when it is absent. This is a matter for listening tests, which indicate that when speech is present the appraisal of noise improvement indeed is affected, but to a small extent. This is shown in Table III.

As a matter of orientation minus ten vu represents about the average volume which reaches the toll switchboard for transmission, while plus eight vu is attained by only one in about 10,000 talkers. It is apparent that over the practical range of talking volumes nearly the full improvement is maintained, even though the noise may rise many decibels during the transmission of a syllable. Suppression of the noise during intervals between syllables seems to be of controlling importance, which probably is related to the relief of the ear from fatigue. That the improvement is no mere illusion of judgment has been shown by articulation tests. It also has been demonstrated in a striking way by an experimental arrangement such that the noise without the compandor not only interfered with but completely obliterated the speech, whereas with the compandor the speech was intelligible.

Circuit Description

Compressor and Expandor

The functional elements of a compandor are shown in Figure 5 as they are arranged at one terminal of a 4-wire circuit. The other end of the circuit is equipped similarly.

The talking path through the compressor consists of a vario-losser followed by an amplifier. The control path works from the output of this amplifier back to the vario-losser. In the control path there is an amplifier followed by a rectifier, the current from which controls the loss of the vario-losser.

Let E_i be the input voltage, and E_o be the output voltage of the compressor. Since the vario-losser consists of a variable resistance R bridged across a circuit of considerably higher resistance, the voltage across R is proportional to the value of R and to the input voltage. Thus $E_o = K_1RE_i$. The relationship desired for a 2/1 compressor is $E_o = KE_i^{1/2}$. This relationship will be met if R is inversely proportional to E_o. If the variable element is a copper oxide varistor, this relationship is approximated. The current-voltage characteristic of copper oxide varistors can be approximated over a considerable range by the expression $I = a\epsilon^{bE}$, where I and E are current and voltage, a and b are constants.

$$\frac{1}{R} = \frac{dI}{dE} = ba\epsilon^{bE}$$

whence

$$R = 1/bI$$

If the control system is linear, so that $I = K_2E_o$, then, as desired,

$$R = 1/K_2E_o$$

In the expandor the talking path also contains a vario-losser followed by an amplifier, but in this case the control amplifier-rectifier is connected ahead of the vario-losser. The latter is a variable resistance R in series with a circuit of considerably lower resistance. Thus the current through the circuit is determined mainly by R. The voltage across the low output impedance is proportional to this current so that $E_o = E_i/K_1R$. As before, the rectifier is linear, and the varistor is copper oxide, so that $R = K_2/E_i$, whence $E_o = KE_i^2$, as desired for a 2/1 expandor.

The cross control consists of another rectifier connected to the expandor control amplifier, with the d-c output connected through a one-way element to the vario-losser of the compressor. Thus at the same time that the expandor loss is being reduced by input, an approximately equal loss is being added to the compressor. The one-way device at the output of the cross control, a copper oxide varistor, makes the action unilateral.

Figure 8. Front view of compressor (cover removed), Western Electric type 1*A*

MATCHING CHARACTERISTICS

The compressor and the expandor should be exactly complementary, so that the over-all circuit will have a constant loss at all amplitudes. It was a design objective to keep deviations from this relation to less than one decibel for all causes, including manufacturing tolerances, aging, temperature, and power supply variations, with the added condition that any compressor must work with any expandor.

The variations contributed by amplifiers are made negligible by designing both line and control amplifiers with about 35 decibels of negative feedback, which reduces any individual variation by a factor of 50 or more.

As ordinarily used, copper oxide varistors are subject to relatively large variations because of manufacturing tolerances, aging, and temperature changes. As used here, the copper oxide disks are superaged in the factory. It is estimated that this produces an aging effect equivalent to about 75 years of ordinary service, and that further changes will be of the order of three per cent in resistance over a 20-year period. After this process, the disks are selected by individual tests, high and low resistance disks being combined so that the four arms of the vario-lossers are balanced.

To minimize variations caused by ambient temperature changes, the varistors are mounted in a small oven, the temperature of which is controlled by a thermostat to an accuracy of about plus or minus three degrees Fahrenheit.

Because of the design precautions taken, the control rectifiers are closely alike; however, their input-output characteristic is not strictly linear. For this reason, the rectifiers are adjusted to work over the same portion of the curve by adjusting their output currents to be equal at a normal testing tone input of zero "dbm" at a zero-level point in the circuit. It is found by experience that varistors differ most in resistance at low control currents. For this reason, an additional adjustment is made by means of a potentiometer at a circuit input corresponding to the bottom of the companding range. This adds a small direct current derived from the 130-volt plate battery to the current produced by the control rectifier, thus making the input-output curve cross the ideal again at this input.

PERFORMANCE

Figure 2, which shows the input-output characteristics of idealized compressors and expandors, contains dashed lines which show typical measured characteristics for comparison. These are matched to a degree such that the over-all circuit loss measured at different input amplitudes will not vary more than plus or minus one decibel, and in the usual case will be in the range plus or minus 0.25 decibel. These figures include all variations except the effect of loss variations in the circuit between compressor and expandor.

Frequency characteristic distortion, modulation products, and noise from apparatus sources all have been reduced to the point of contributing only a negligible amount on a high-grade 200–3,500-cycle message channel.

DETAILED DESCRIPTION

Figures 6 and 7 show the circuits of the compressor and expandor in some detail. The line and control amplifiers each comprise two stages, impedance coupled, with about 35 decibels of negative feedback for gain stabilization. Screwdriver controlled potentiometers in the feed-back path are used for fine adjustment of gain. Heater-type pentode tubes are used throughout.

Figure 9. Front view of expandor (cove removed), Western Electric type 1*A*

The temperature inside the oven may be measured by means of a thermistor disk about the size of a dime, which is inside the potting can holding one of the vario-lossers. This thermistor is a portion of a Wheatstone-bridge circuit, the other arms being made up of resistances in a vacuum-tube cathode circuit. The condition of balance of the bridge, which is an indication of the oven temperature, may be determined by the same testing equipment[5] used to maintain other circuits in the telephone office. The currents flowing in the various cathode circuits and in the control rectifier output are measured also with this same test set by connection to the receptacle "test" on the panel.

In case of a thermostat failure causing permanent closure, damage to the

An Oscillograph for Recording Transient Recovery Voltages

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THE importance of the effects of the magnitude and rate of rise of the transient recovery voltage upon the difficulty of arc interruption in circuit breakers has been recognized for a number of years. It is known that differences in the interrupting performance of circuit breakers may be obtained at different locations in power systems, even though the power-frequency voltages and short-circuit currents involved may be similar.[1] In general, the greater the rate at which the transient recovery voltage builds up and the greater the final value that it attains across the arc space as the arc extinction process occurs, the more difficult it is for the circuit breaker to clear the circuit. There is not, however, any simple quantitative relationship between these transient values and the arcing time of a circuit breaker or its interrupting capacity limit. Individual short-circuit interruptions with a given circuit breaker and circuit conditions may show considerable variations in arcing time. For such and other reasons the added complication of transient recovery-voltage rate and magnitude limits has not been added to circuit breaker interrupting capacity ratings or specifications.[2]

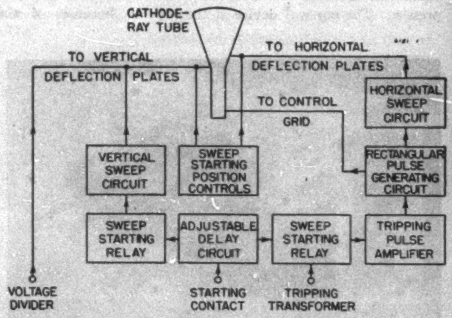

Figure 1. Block diagram of main circuits in oscillograph

Interrupting capacity ratings applied by manufacturers to circuit breakers generally are intended to be sufficiently conservative to hold under the most severe recovery conditions likely to be encountered in practice. In the larger ratings, test verification of this assumption may be difficult or impossible to obtain. In certain applications full rating short circuits result in low recovery-voltage rates. Such a condition well might be considered in specifying interrupting capacity requirements, especially where the choice otherwise lies between adjacent standard ratings.

Although the transient recovery characteristics of circuits in which circuit breakers are connected may be calculated,[3,4] the numerical results desired may be obtained in many cases only after the expenditure of considerable effort and the making of a number of simplifying assumptions. Indirect tests at low voltages may be made to deter-

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varistors is prevented by a temperature operated fuse which opens the heater circuit when the oven temperature becomes greater than 140 degrees Fahrenheit.

The pads "out" (Figure 6) and "in" (Figure 7) are the adjustment pads P_C and P_E previously discussed. They are arranged to plug into tube sockets, and may be selected in 0.25-decibel steps from 0 to 20 decibels. The losses of the two pads in one direction of transmission always must sum up to 20 decibels to maintain complementary action in this compandor.

Physical Description

The compressor and the expandor, which are shown in Figures 8 and 9, are 7-by-19-inch panels designed to mount on relay racks. Stud-mounted components are on the rear while wiring and vacuum tubes are on the front of the panels. Front covers have cutouts for access to testing jacks and controls. Each panel weighs approximately 25 pounds. Shop-wired bays normally include 12 units, which may be compressors, expandors or a combination of both. Patching and monitoring jacks, a 4-wire monitoring and talking panel, and miscellaneous trunks as required also are provided in the bay.

The power supplies required are 130 volts for plates, 24 volts for filaments, and 24 volts for the ovens. A power panel is also available so that the filament circuit and oven heaters may be operated from an a-c source, if desired. It is a $10\frac{1}{2}$-inch panel with sufficient capacity to supply 12 units. The total power required is about 40 watts for each unit.

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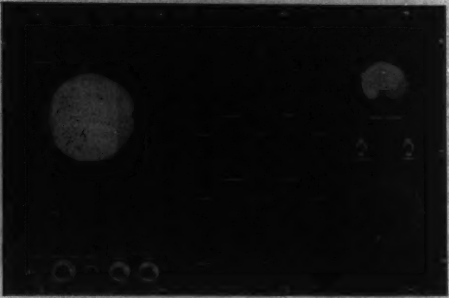

Figure 2 (left). Front panel of oscillograph

Figure 3 (below). Oscillograph with camera and tripping transformer

mine transient performance of circuits viewed from the circuit breaker location.[5,6] Such tests require that the circuits involved be de-energized during the test. They do not indicate any modifications of the recovery-voltage transient resulting from circuit breaker arc voltage or residual arc conductivity after the main current zero. Where interruption tests are made, either in the field or in the high-power test laboratory, there is considerable interest and satisfaction in directly measuring these voltages. Such measurements lie within the field of application of the moderate or high-speed recording cathode ray oscillograph.

Methods of Measurement

A requirement peculiar to these recovery-voltage measurements involves the ability to record transient voltage variations occurring in a few microseconds, and maintaining this high recording speed sensitivity over a time interval of several cycles of power frequency. Two basically different methods have been employed for timing the record on the photographic film and obtaining a sufficiently expanded time scale.

One method utilizes a moving film oscillograph.[7,8] The film is attached to a drum which is rotated at speeds up to several thousand revolutions per minute. The cathode ray beam is deflected by the variations of voltage across the circuit breaker along an axis at right angles to that of the film travel. The beam is energized for a period of time sufficient to include the interruption process. During this exposure time the film may complete a number of revolutions with the result that a similar number of overlapping traces appears on the record. The separate traces are usually distinguishable, and since the total length of trace on the film is several feet, low and moderate transient recovery-voltage frequencies and rates of rise can be resolved satisfactorily. With this method the complete history of the arc voltage in the circuit breaker from the beginning of the arc to its final extinction is available on the oscillogram. For most purposes, however, the important parts of the record are only those regions in the immediate vicinity of the current zeros, wherein the transient recovery voltages appear. The moving-film oscillograph is electrically relatively simple, and timing the period of recording is easily accomplished. There are, nevertheless, a number of disadvantages. The maximum speed at which the film may be driven is limited by mechanical considerations to a value in the order of 0.005 centimeter per microsecond. This is not sufficient for easy or accurate resolution of the higher frequency components that may exist under some test conditions. For example, one cycle of a 100,000-cycle wave would appear 0.5 millimeter long at this speed. The high-speed revolving drum with its necessary accurate balancing, film supporting means, and motor drive involve mechanical complications not present in oscillographs using a stationary film.

In the other method of recording transient recovery voltages the cathode-ray beam is deflected in a direction perpendicular to that caused by the voltage being measured to give the time axis.[1,9] The oscillogram is produced on stationary film. The total deflection distance in the time axis direction is limited to a few inches in this case. Without difficulty the speed of deflection may be made much greater than obtainable with mechanically driven film. A relatively intensive examination of a small fraction of the total interruption process is thus possible. The main difficulty encountered is that of timing the start of this high speed short duration time sweep to occur at the proper instant to include the transient recovery voltage at the end of the last half cycle of arcing. Probably the most satisfactory solution so far employed for this problem involves the use of a saturating-core current transformer to produce a voltage peak in the secondary winding at approximately current zero in the primary. In another case,[10] the beam is swept back and forth by a sine-wave power-frequency deflection voltage of much greater peak value than necessary to move the beam across the fluorescent screen width. The timing is arranged so that the arc extinction period occurs during the time the beam is on the screen. A number of sweeps appear on the same part of the record, and the right and left-hand travel is somewhat difficult to follow.

Description of Oscillograph

The oscillograph described in this paper utilizes a high-voltage sealed-off cathode-ray tube of the same kind as used for impulse testing.[11] The record is produced on a stationary film. The

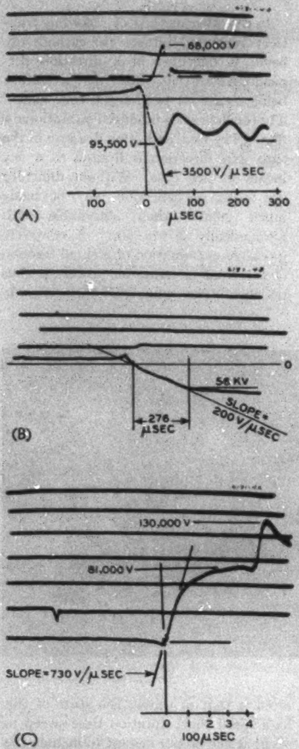

Figure 4. Recovery voltage oscillograms taken on 110-kv line-to-ground short-circuit tests at Newark substation of Pacific Gas and Electric Company

A. Test at 8,290 amperes. The 60-cycle crest recovery volts = 88,000

B. Test at 14,400 amperes

C. Test at 5,200 amperes

necessary high time-sweep speed is obtained by moving the beam horizontally across the screen in a short period of time in the region of current zero. This sweep is repeated once each half cycle from a time definitely before the arcing contacts of the circuit breaker separate until the final arc interruption. The successive horizontal sweeps are separated by superimposing them on a relatively slow uniform downward sweep or drift of several cycles duration. The time interval between the horizontal sweeps, when the screen is dark, is nearly a half cycle. The complete record thus appears as a series of spaced horizontal lines with vertical deviations produced by the voltage across the circuit breaker. The last and lowest line contains the recovery-voltage transient at the final current zero.

This method of operation permits use of a high intensity beam with about the same time integral of beam power to the fluorescent screen per horizontal sweep as when the tube is used in oscillographs for impulse testing. This is possible because the actual fraction of the time the beam is energized is very small and because in the successive sweeps the beam does not retrace the same screen area. Because the film is stationary the usual medium-persistence screen may be used. The advantages of a simple fixed camera are apparent. The camera used in this case allows simultaneous visual observation and photographic recording of the traces on the screen. This opportunity of visual observation has been found very useful in giving immediate information about the duration of arcing, the polarity at interruption, the shape of the recovery voltage transient and the effect of the residual arc conductivity in modifying it.

Although the circuits contained in this instrument are more complex than necessary with the moving-film type of oscillograph, most of the extra equipment is made up of easily available radio receiving tubes and parts. The oscillograph and auxiliary equipment are sufficiently compact and portable to be transported easily to test locations in a passenger automobile.

A block diagram of the main component circuits is shown in Figure 1. The cathode-ray tube used is an RCA-912. This tube may be operated with a total beam accelerating voltage of 15,000 volts. It was decided in this case to use a maximum value of 10,000 volts since the recording speeds necessary are considerably below the tube limit. The lower voltage power supply is lighter in weight and more compact.

The sequence of operation may be summarized as follows:

1. Initiation of the vertical and horizontal sweep actions depends upon the closing of an external starting contact a short time (a few cycles) before the circuit breaker arcing contacts separate. This contact should remain closed during a portion of the opening stroke of the circuit breaker sufficient to include the interruption process.

2. Actual initiation of the afore-mentioned sweeps follows the closing of two small fast-acting relays within the oscillograph. To aid in timing the starting of the two sweeps, an adjustable delay circuit is interposed between these relays and the external starting contact. The vertical sweep starts immediately upon the closing of its starting relay. On closing the horizontal sweep-starting relay simply completes the circuit from the externally mounted tripping transformer to the horizontal sweep voltage generating circuit. This tripping transformer produces a sharp peak of voltage once each half cycle of the test current at a few degrees (adjustable) before current zero. The successive horizontal sweeps are initiated by these voltage pulses and are thus properly timed to include the current zero region.

3. The control grid of the cathode ray tube is normally maintained at a bias voltage sufficiently negative to prevent the formation of an electron beam. For the short duration of the individual horizontal sweep, however, the beam is released by the application of a rectangular positive voltage pulse from the horizontal sweep circuit.

4. The recovery voltage appearing across the test circuit breaker is reduced to a lower and accurately proportional value by means of a capacitance-type voltage divider. This voltage is fed through a shielded transmission line from the circuit breaker location to the oscillograph deflection plates.

Figures 2 and 3 show the oscillograph, tripping transformer, and camera. Controls brought out to the front panel of the instrument are for adjusting the starting position of the beam at the beginning of the sweeps, the speeds of both the vertical and horizontal sweeps, the sweep-starting time delay, the cathode-ray tube beam focus and intensity, and the total beam accelerating voltage. This last quantity is indicated by a meter on the panel. A push button for producing a test sweep also is provided. Input terminals shown at the lower left of the panel are for the 115-volt 60-cycle power source, the secondary of the tripping

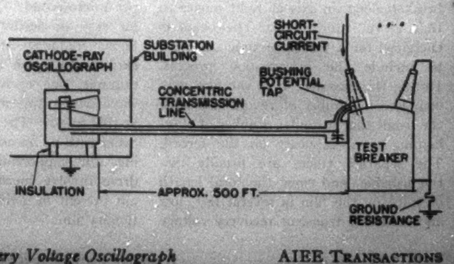

Figure 5. Connection of cathode-ray oscillograph to test circuit breaker for Newark tests

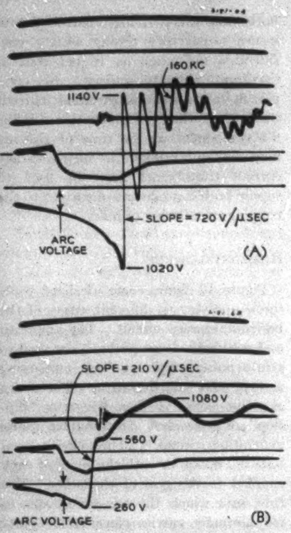

Figure 6. Recovery voltage oscillograms taken on short-circuit tests with laboratory generator at Stanford University

A. Recovery showing high frequency oscillation

B. Recovery showing damping due to arc-space conductivity

transformer, a connection for the external sweep-starting contact, and one concentric cable terminal for each vertical deflection plate.

Test Results

This oscillograph has been used on interruption tests on power systems and as a key instrument in laboratory studies of arc-extinction phenomena.

Figure 4 shows oscillograms taken during short-circuit tests made at the Newark substation of the Pacific Gas and Electric Company, July 8 and 15, 1945. The single-pole test circuit breaker was connected from a section of the 110-kv bus to ground. Figure 4A is the oscillogram obtained for a current of 8,290 amperes rms, corresponding to an equivalent 3-phase interruption of 1,600,000 kva. The current in this case was limited principally by reactance of transformers within the substation grounds. The rate of rise of recovery voltage obtained was high. It is interesting to note the restriking voltage built up across the arc space on the previous half cycle. In another test, Figure 4B, the interrupted current was 14,400 amperes and the equivalent 3-phase interruption was 2,720,000 kva. In this test the current was limited mainly by the impedance of several long transmission lines feeding into the substation. The voltage at the receiving end of these lines was pulled down to a low value during the flow of current. A relatively low recovery rate was obtained, as a result of the time required for restoration of normal voltage on these lines after the current ceased to flow. Figure 4C was obtained for a test on a smaller circuit breaker. The test current was limited to 5,200 amperes with reactance at the substation. In order to reduce the severity of the recovery-voltage conditions, an open-ended 110-kv transmission line was connected in parallel with the circuit breaker. The length of this line was about 40 miles. The oscillogram shows the returning wave front reflected from the open end of the line.

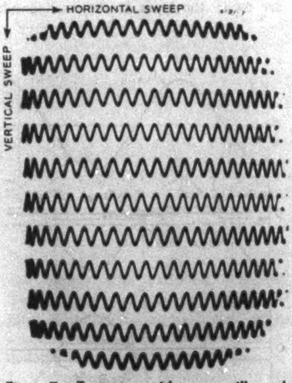

Figure 7. Test sweep with same oscillograph adjustments as for Figure 6

200-kc timing wave

Figure 5 shows the connections between the test circuit breaker and the cathode-ray oscillograph used in the foregoing tests. Since the short circuits were line to ground there was possibility of a considerable difference of potential between the ground connection at the location of the circuit breaker and grounds at other parts of the substation. For reasons of safety and convenience all measurement and control equipment was centered in the substation building about 500 feet from the test circuit breaker. A connection to the capacitance tap in the line-side bushing of the circuit breaker was brought through a concentric transmission line to one of the vertical deflection plates of the cathode-ray oscillograph. The outer conductor of this line was connected solidly to the circuit breaker tank. The oscillograph housing and the other vertical deflection plate were connected to the other end of the outer conductor. A capacitive division of the voltage across the circuit breaker was obtained according to the ratio of the capacitance between the bushing tap and the bushing center conductor to the capacitance of the transmission line center conductor, the bushing tap, the deflection plate circuit, and an auxiliary shunting capacitor to ground. The cathode-ray oscillograph was mounted on insulating supports and the 115-volt 60-cycle power required for its operation supplied through a low capacitance insulating transformer. In this way the potential of the oscillograph housing was allowed to follow freely that of the circuit breaker ground and only the difference of voltage across the circuit breaker caused deflection of the beam. Actually, the voltage appearing between the oscillograph housing and the ground connection near by was only about 200 volts during the test current flow of 14,000 amperes.

Two oscillograms taken in laboratory

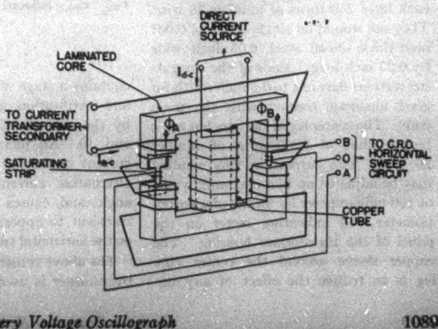

Figure 8. Diagram of core and windings of tripping transformer

short-circuit tests of a small circuit breaker are shown in Figure 6. The circuit voltage was 1,300 volts and the interrupted current 600 amperes. The oscillogram in Figure 6A shows a large high-frequency oscillation with relatively low damping. Figure 6B was obtained for another test interruption several minutes later under identical circuit conditions. In this case the effect of the arc conductivity during the extinction period is indicated by the complete damping of the high-frequency oscillation and the reduced rate of rise of the recovery-voltage transient. Simultaneous records of the current in the arc space during this critical period of time show for the first test practically no current flow after the normal zero, but a considerable amount, in the order of two amperes, for a short time in the second test. Results of these current measurement tests will be reported at a later date.

The oscillograms of Figure 6 were taken with a fast horizontal sweep. They represent periods of approximately 2 degrees of the 60-cycle power-frequency wave. Figure 7 shows a 200-kc timing wave test with the same oscillograph adjustments as used for Figure 6. The horizontal sweeps were tripped by passing a steady 60-cycle current through the primary of the tripping transformer.

Circuit Details

TRIPPING TRANSFORMER

The principle of operation of the saturating-core peaking or horizontal-sweep tripping transformer is illustrated by Figures 8 and 9. The primary winding consists of four coils, two on each outside leg of the core. Each coil has 22 turns of number 14 wire. The coils may be connected in series, series-parallel, or parallel. The operating range is approximately 3 to 30 amperes per coil. The impedance of the primary winding with the series connection is about 0.5 ohm. The secondary windings each have 250 turns of number 38 wire. They are wound on single strips of 0.007 inch thick silicon steel, 0.05 inch wide by 0.25 inch long. Ends of the saturating sections flare out to 0.5 inch width for good magnetic contact with the main core. The center leg of the core has a d-c bias winding of 1,200 turns of number 26 wire. The current in this winding may be adjusted up to a maximum value of 100 milliamperes by means of a potentiometer and indicating meter on the panel of the transformer housing. The copper sleeve around the center core leg is to reduce the effect of any unbalance of the a-c magnetomotive force upon the d-c winding circuit.

Figure 9 shows wave forms, somewhat idealized, applying to the tripping transformer. The result desired is to produce sharp pulses of voltage at a controllable small angle before the current zeros in the primary winding. The net fluxes Φ_A and Φ_B in the outside core legs may be considered as the sums of component fluxes resulting from the a-c primary and d-c bias magnetomotive forces. The magnetic strips through the secondary windings are saturated except during short periods in the vicinities of the next flux reversals. The corresponding induced voltage peaks have a maximum value in the order of two volts and a width of about two degrees at half amplitude. Only the negative pulses are used; the input to the oscillograph contains a stage of voltage amplification and rectification, with a result indicated by the peaked voltages e_R in Figures 9 and 10. An increase in the d-c bias current causes the peaks e_R to lead the alternating current zeros by a larger angle and causes the recovery-voltage transient to appear farther to the right on the horizontal sweep.

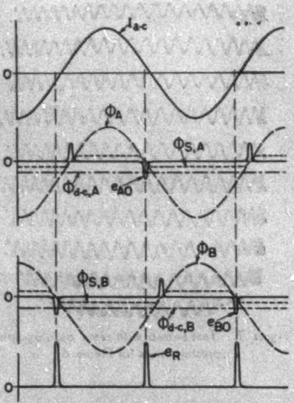

Figure 9. Idealized wave forms in tripping transformer

Φ_A, Φ_B, net fluxes in outside core legs

$\Phi_{d\text{-}c,A}$ $\Phi_{d\text{-}c,B}$, d-c components of flux in outside core legs

$\Phi_{S,A}$, $\Phi_{S,B}$ fluxes in saturating strips

e_{AO}, e_{BO}, induced voltages in secondary windings

The above construction of the tripping transformer is used not only to provide horizontal sweep timing control, but to insure satisfactory timing with a considerable (as much as 10 per cent at maximum primary current) d-c component in the main short-circuit current wave at the time of interruption. If the d-c component at the time of clearing may be appreciable, a high-accuracy current transformer must be used to supply both components of current to the tripping transformer primary.

HORIZONTAL SWEEP

Figure 10 shows some idealized wave forms applying to different stages of the horizontal sweep circuit. The amplified and rectified voltage peaks e_R are used to trip a pulse-sharpening stage employing a type 884 gas thyratron. The sharp negative peaks *A* then trigger a "flip-flop" or "one-shot" rectangular pulse generating circuit.[13] This circuit consists of two tubes with the plate of each coupled to the grid of the other. The tube into which the triggering pulse is fed normally carries considerable plate current at a low plate voltage. The other tube is biased beyond cutoff. Upon triggering, the plate voltage of the first tube rises quickly to full plate supply value and remains there for a period of time during which the second tube carries current. The duration of this high plate voltage ends abruptly when the circuit restores itself to the original condition. The length of the practically rectangular pulse of plate voltage *B* on the first tube is mainly a function of the time constants of the grid-plate coupling circuits.

The control grid of the cathode-ray tube normally has a sufficiently negative voltage with respect to the cathode to prevent the formation of an electron beam. It is coupled through a capacitor to the rectangular pulse generating circuit, so that the grid potential rises in the positive direction with the rectangular pulse. The amplitude of this change of grid voltage *D*, about 150 volts, allows the production of a strong beam current *C*. The actual beam intensity is adjusted by controlling the original negative bias on the control grid.

The rectangular pulse of voltage *B* also is applied to a resistance and capacitance in series. The resistance is high enough so that the capacitance is charged to but a small fraction of the amplitude of the rectangular pulse at the end of the pulse. The rate of change of the voltage *E* across the capacitance is very nearly constant under these conditions. The voltage *E* is supplied to a push-pull voltage amplifier, the output of which is connected to the horizontal deflection plates of the cathode-ray tube to produce a nearly linear horizontal time-sweep. Any one of four different sweep speeds may be selected by means of a multi-contact switch operated from the front panel. The times available for a complete horizontal sweep are approximately 20, 10, 5, and 2 degrees on a 60-cycle basis.

Vertical Sweep

Two 0.05 microfarad capacitances, one charged to +1,250 volts and the other to −1,250 volts with respect to ground, are discharged in series through a type 6AC7 pentode upon the closing of the sweep-starting relay. The constant-current characteristic of the pentode produces a nearly uniform rate of change of voltage across these capacitances. This voltage is supplied to the vertical deflection plates through 2-megohm resistances in each lead. Voltages to be measured are fed to the vertical deflection plates through 0.015 microfarad capacitances. The rate of change of vertical sweep voltage is low enough to change the charge on the 0.015 microfarad capacitances without much voltage drop in the 2-megohm resistances. Though the continuous voltage rating of the 6AC7 pentode is exceeded greatly in this circuit, the duration of application is so short that no trouble has been experienced over more than a year of frequent use. Different vertical-sweep speeds are obtained by varying the cathode bias resistance for the 6AC7 pentode.

Other Circuits

Four rectified d-c power supplies are built into this oscillograph. Accelerating and focusing voltages for the cathode-ray tube are obtained from a voltage-doubling circuit using type 2X2 rectifier tubes. Voltage for the vertical sweep circuit and for the last amplifier stage of the horizontal sweep circuit is obtained from a 2,500-volt supply with midpoint grounded. Two conventional 300-volt rectifier and filter combinations are used, one for the amplifier circuits, and the other for the sweep-starting relay circuit. These relays are operated by current flow in the plate circuit of a type 884 thyratron. The tripping grid voltage for this thyratron is developed across a capacitance which is charged through an adjustable resistance in series with the external sweep-starting contact. Steps of this charging resistance controlled from the front panel may be set to cause delays of 0 to 6 cycles in 1-cycle intervals.

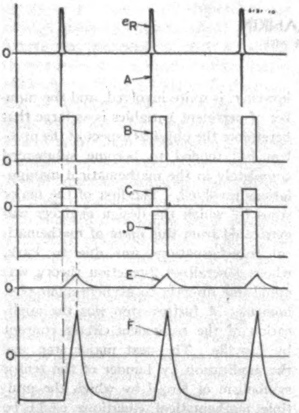

Figure 10. Idealized wave forms in horizontal sweep circuit

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Asynchronous and Single-Phase Operation of Synchronous Machines

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ASYNCHRONOUS operation of synchronous machines has been the subject of numerous articles in the technical literature, for example, the articles of Fechheimer,[1] Hay and Mowdawalla,[2] Putman,[3] Park,[4] Linville,[5] Lauder[6]. The fact that articles on this subject continue to be of primary interest to design and operating engineers is attributable to the importance of satisfactory operation at asynchronous speeds. It is only through a precise knowledge of the conditions present in asynchronous operation that machines can be designed for and operated through these asynchronous periods satisfactorily. As examples of the extreme conditions which can arise during asynchronous operation, it should be noted that stalling a synchronous machine with the field voltage applied can result in pulsating torques greater than those obtained by inadvertent synchronizing 180° out of phase;[6] or that the currents in the damper bars on the trailing half of a pole can be considerably greater than the currents in the corresponding bars on the leading half of the pole, and the resulting unequal heating will influence the design of the damper winding to a considerable degree. Because of the importance of these and similar phenomena, the problem of asynchronous operation must be studied continuously until a complete, easily visualized, and easily applied solution is obtained.

The analysis of asynchronous operation, however, is quite involved, and the number of pertinent variables is so large that heretofore the physical aspect of the problem has tended to become submerged completely in the mathematical manipulations involved. The first of the major steps by which the design engineer was extricated from this maze of mathematical transformations was due to Park, whose generalized 2-reaction theory was applicable directly to asynchronous phenomena. A further step was the application of the equivalent-circuit concept by Linville. The next major step was the application by Lauder of the tensor symbolism of Kron[7] by which the multiple mathematical equations could be handled by groups rather than individually.

Important as these latter steps were, however, the calculations were still too time-consuming and labyrinthine to permit their application to routine design problems. The most promising solution to these difficulties appears to be the generalized equivalent circuits of Kron.[7,8,9] With these circuits the problem is reduced to the determination of currents and voltages in relatively simple networks. This not only eliminates the distracting mathematical transformations heretofore employed but also presents an easily visualized method in which the physical characteristics retain their importance and are not submerged in the mathematics.

It is accordingly the purpose of this article to present an analysis of the asynchronous and single-phase operation of a synchronous machine based on the generalized equivalent circuits of Kron. The fundamental equations of Park will be used as the starting point, but the succeeding mathematical operations will be guided by the aforementioned tensor publications of Kron. The asynchronous torque and current relations will be presented in terms of the currents and voltages of simple equivalent circuits, and, in addition, a numerical comparison will be given of the asynchronous and single-phase operating characteristics of synchronous machines with complete and incomplete end rings on the damper windings. This comparison has direct value in design engineering not only as a study of the various torques, but also as a picture of the distribution of the damper-bar currents in the leading and trailing pole halves.

The methods here presented are intended for use with the a-c network analyzer,[10] and all the torque and current expressions are given directly in terms of current and voltage readings obtained from the network analyzer.

This paper is divided into the following four main parts:

A. Derivation of the equivalent circuits for constant-speed asynchronous operation.

B. Numerical example of the application of the equivalent circuits of *A*.

C. Derivation of the equivalent circuits for constant-speed single-phase operation.

D. Numerical example of the application of the equivalent circuits of *C*.

This article is based upon three recent papers by the author,[11-13] which were published to provide a sound basis for the present study. It will be assumed that the reader is familiar with their contents.

The equivalent circuits for asynchronous operation are based on the circuits of reference 8, while the equivalent circuit for single-phase operation is based

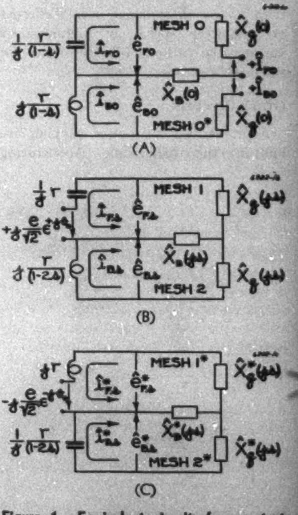

Figure 1. Equivalent circuits for constant-speed asynchronous operation

A. Rotor excitation only
B. Stator excitation only
C. Stator excitation only

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The author acknowledges the co-operation and contributions of his associates in the General Electric Company in the study presented in this paper, S. B. Crary for his suggestions and constructive criticisms, C. E. Kilbourne for his stimulating interest and for placing at the author's disposal several unpublished studies of asynchronous operation by the members of his division, and C. Concordia for his contributions to the single-phase study. Particular appreciation and thanks are extended to Gabriel Kron whose published and unpublished papers on tensor analysis and equivalent circuits guided the mathematical analysis presented here.

on the circuits of the primitive machine presented in reference 9.

Results

Figure 1 gives the equivalent circuits (spin circuits) from which the asynchronous operating characteristics of a synchronous machine may be determined by meter readings of the currents, voltages, and powers. The formulas by which the machine characteristics are expressed in terms of the spin-circuit quantities are given in part *A*.

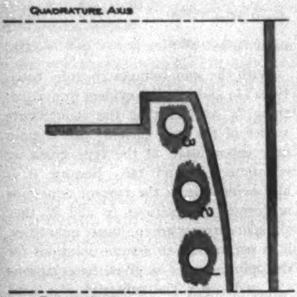

Figure 2A. Numbering of damper bars

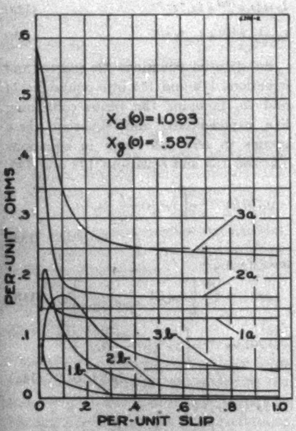

Figure 2B. Impedance functions $\hat{x}_d(js)$ and $\hat{x}_q(js)$ for the synchronous machines used in the numerical examples

$\hat{x}(js) = x - jR$ with x and R given by these curves

a—real component (x)

b—imaginary component (R)

1—direct-axis quantities; complete and incomplete end rings

2—quadrature-axis quantities; complete end rings

3—quadrature-axis quantities; incomplete end rings

As an example, $\hat{x}_q(js)$ for the machine with complete end rings at 0.5 slip is $0.170 - j0.024$

A comparison of the calculated operating characteristics of synchronous machines with complete and incomplete end rings is given by Figures 2A–15. These figures are discussed in part *B*. From these curves it is evident that by means of the spin circuits, a solution to the problem of asynchronous operation at constant speed has been obtained which, for ease of application, elimination of mathematical transformations, and completeness of treatment, surpasses the solutions previously published for this problem. The numerical example itself possesses immediate practical value. The curves showing current distribution in the damper windings present a picture of the unequal heating which the damper winding must be designed to handle satisfactorily, while the torque curves illustrate the dangerous torque pulsations which exist not only at stalling speeds but throughout the entire speed spectrum.

Figure 17 gives the equivalent circuit from which the single-phase operating characteristics of a synchronous machine may be determined by meter measurements of the various currents, voltages, and powers. The type of single-phase loading is shown in Figure 16. The formulas which express the machine operating characteristics in terms of the spin-circuit quantities are given in part *C*. Note that this spin circuit permits the direct determination of the harmonic currents and torques to any harmonic order.

A numerical example of the calculated single-phase operating characteristics of the synchronous machines described in part *B* is given by part *D*. It is shown there that even in single-phase operation, a considerable difference exists in the loading of the leading and trailing damper bars. In the case of the machine with incomplete end rings on a pure resistance load, the 240-cycle currents and torques are of the order of ten per cent of the corresponding 120-cycle quantities.

Unit Currents

All currents in the following study are expressed in per unit. Such dimensions are adequate in studying machine operation, but for design and manufacture the actual ampere values must be known. That the stator currents are in per unit of the rated stator current is obvious, but the unit values of the rotor-circuit currents are not at all obvious. In references 11 and 12, the author points out that unit rotor-circuit currents are entirely arbitrary, but that after they are selected the per-unit rotor-circuit impedances must be calculated accordingly. In this article, the x_{ad} bases of reference 12 are used.

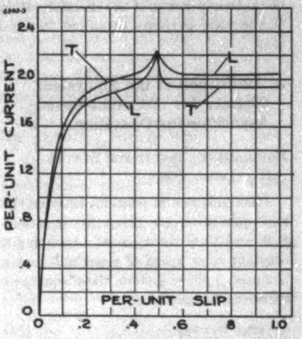

Figure 3. Current in damper bar 1 of machine with complete end rings

T—Trailing

L—Leading

Unit field current is that current which at synchronous speed will induce in each stator phase a voltage $x_{ad}i_{ao}$. This current is expressed in amperes by equation 35*a* in terms of i_{ao} and the machine design constants. Unit current for the additional rotor circuits, both direct and quadrature axes, is that current which when flowing in a damper-bar circuit of 100 per cent pitch will induce in each stator phase, at synchronous speed, a voltage $x_{ad}i_{ao}$. This current is expressed in amperes by equation 35*b* in terms of i_{ao} and the machine design constants.

$$\frac{I_{fdo}}{(^3/_2)\, i_{ao}} = \frac{4}{\pi}\frac{A_{d1}}{F_{d1}}\frac{N}{K_pK_dP_1}\frac{1}{N_{fd}} \text{ (peak)} \quad (35a)$$

$$\frac{I_{xdo}}{(^3/_2)\, i_{ao}} = \frac{4}{\pi}\frac{A_{d1}}{D_{d1x}}\frac{N}{K_pK_dP_1}\frac{1}{N_{xd}} \text{ (peak)} \quad (35b)$$

All impedances are calculated with the base-current ratios, equations 35*a* and 35*b*. Formulas for the impedances, using these base-current ratios, were derived in reference 13.

The various current expressions are in per unit instantaneous rather than effective values, but these instantaneous values are in per unit of the unit currents,[11] equations 35*a* and 35*b*. Hence, when it is desired to determine the effective value of a slip-frequency current, it must be taken in per unit of the effective values of the base currents equations 35*a* and 35*b*.

A. Derivation of the Equivalent Circuits for Constant-Speed Asynchronous Operation

Equivalent Circuit for Stator Excitation Only

The fundamental relations between the axis quantities are given by equations 8, 9, and 10 of reference 4. Assuming a constant rotor speed of v per unit and a balanced 3-phase system, these equations reduce to the following:

$$e_d = p\psi_d - ri_d - v\psi_q \tag{1a}$$

$$e_q = p\psi_q - ri_q + v\psi_d \tag{1b}$$

Let the following complex spin quantities be defined.

$$\hat{\psi}_F = \frac{\psi_d + j\psi_q}{\sqrt{2}} \tag{2a}$$

$$\hat{\psi}_B = \frac{\psi_d - j\psi_q}{\sqrt{2}} \tag{2b}$$

Let the spin quantities $\hat{e}_F$ and $\hat{e}_B$, $\hat{i}_F$ and $\hat{i}_B$, be defined analogously in terms of their corresponding axis quantities. Since the axis quantities are real, it is evident that the spin quantities are conjugate complex quantities.

Since all the rotor-circuit voltages are assumed zero, the axis linkages are given by the following operational functions of the axis currents. The first of these relations, equation 3*a*, was given by equation 8*a* of reference 13, and the second is obtained in an analogous manner.

$$\psi_d = -x_d(p)i_d \tag{3a}$$

$$\psi_q = -x_q(p)i_q \tag{3b}$$

The spin linkages are obtained in terms of the spin currents by substituting equations 3*a* and 3*b* into equations 2*a* and 2*b*. Equations 4*a*, 4*b*, 5*a*, and 5*b* thereby are obtained.

$$\hat{\psi}_F = -x_F(p)\hat{i}_F - x_B(p)\hat{i}_B \tag{4a}$$

$$\hat{\psi}_B = -x_B(p)\hat{i}_F - x_F(p)\hat{i}_B \tag{4b}$$

$$x_F(p) = \frac{x_d(p) + x_q(p)}{2} \tag{5a}$$

$$x_B(p) = \frac{x_d(p) - x_q(p)}{2} \tag{5b}$$

Substituting the spin equations of the type given by equations 2*a* and 2*b* back into the axis equations 1*a* and 1*b* will give the fundamental spin relation of equations 6*a* and 6*b*.

$$\hat{e}_F = (p + jv)\hat{\psi}_F - r\hat{i}_F \tag{6a}$$

$$\hat{e}_B = (p - jv)\hat{\psi}_B - r\hat{i}_B \tag{6b}$$

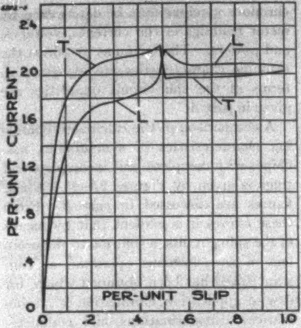

Figure 4. Current in damper bar 2 of machine with complete end rings

T—Trailing
L—Leading

Substituting equations 4*a* and 4*b* into equations 6*a* and 6*b* gives the operational equations 7*a* and 7*b*.

$$-\hat{e}_F = (p + jv)[x_F(p)\hat{i}_F + x_B(p)\hat{i}_B] + r\hat{i}_F \tag{7a}$$

$$-\hat{e}_B = (p - jv)[x_B(p)\hat{i}_F + x_F(p)\hat{i}_B] + r\hat{i}_B \tag{7b}$$

Assuming that balanced 3-phase voltages are applied with the machine running at a constant slip, the phase voltages are as given by equations 8*a*–8*c*; the angular position is as given by equation 9; and the axis voltages are as given by equations 10*a* and 10*b*. These latter voltages are obtained by means of equation 6 of reference 4.

$$e_a = e\cos t \tag{8a}$$

$$e_b = e\cos(t - 120) \tag{8b}$$

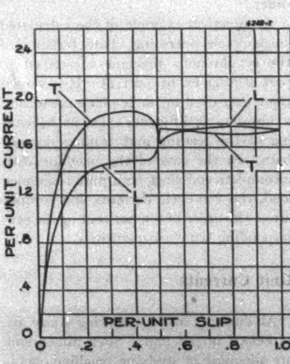

Figure 5. Current in damper bar 3 of machine with complete end rings

T—Trailing
L—Leading

$$e_c = e\cos(t + 120) \tag{8c}$$

$$\theta = (1 - s)t - \theta_0 \tag{9}$$

$$e_d = e\cos(st + \theta_0) \tag{10a}$$

$$e_q = e\sin(st + \theta_0) \tag{10b}$$

Substituting equations 10*a* and 10*b* into the voltage equations analogous to equations 2*a* and 2*b* gives the spin voltages equations 11*a* and 11*b*.

$$\hat{e}_F = 0.707e\epsilon^{+j(st+\theta_0)} \tag{11a}$$

$$\hat{e}_B = 0.707e\epsilon^{-j(st+\theta_0)} \tag{11b}$$

With the spin voltages given by equations 11*a* and 11*b*, it is evident from equations 7*a* and 7*b* that the steady-state values of the spin currents can contain only exponentials of the type given by equations 11*a* and 11*b*. Because it is also evident from the current equations analogous to equations 2*a* and 2*b* that the spin currents are conjugate quantities, it is permissible to assume solutions for the spin currents as given by equations 12*a* and 12*b*. The subscript s is used here to identify these spin currents as belonging to the solution for constant-speed asynchronous (slip) operation.

$$\hat{i}_F = \hat{i}_{F_s}\epsilon^{+jst} + \hat{i}_{B_s}{}^*\epsilon^{-jst} \tag{12a}$$

$$\hat{i}_B = \hat{i}_{B_s}\epsilon^{+jst} + \hat{i}_{F_s}{}^*\epsilon^{-jst} \tag{12b}$$

Substituting equations 12*a* and 12*b* and equations 11*a* and 11*b* into equations 7*a* and 7*b*, grouping the terms containing like exponentials, and equating these groups to zero will result in equations 13*a* and 13*b* and equations 14*a* and 14*b*.

$$+0.707je\epsilon^{+j\theta_0} = \left[\mathfrak{x}_F(js) + \frac{r}{j}\right]\hat{i}_{F_s} + \mathfrak{x}_B(js)\hat{i}_{B_s} \tag{13a}$$

$$0 = \mathfrak{x}_B(js)\hat{i}_{F_s} + \left[\mathfrak{x}_F(js) - \frac{r}{j(1-2s)}\right]\hat{i}_{B_s} \tag{13b}$$

$$-0.707je\epsilon^{-j\theta_0} = \left[\mathfrak{x}_F{}^*(js) - \frac{r}{j}\right]\hat{i}_{F_s}{}^* + \mathfrak{x}_B{}^*(js)\hat{i}_{B_s}{}^* \tag{14a}$$

$$0 = \mathfrak{x}_B{}^*(js)\hat{i}_{F_s}{}^* + \left[\mathfrak{x}_F{}^*(js) + \frac{r}{j(1-2s)}\right]\hat{i}_{B_s}{}^* \tag{14b}$$

Equations 13*a* and 13*b* can be solved electrically by means of the a-c network of Figure 1B; equations 14*a* and 14*b* can be solved similarly by the a-c network of Figure 1C. It is evident that the circuit of Figure 1C is the conjugate of the circuit of Figure 1B, and that the currents and voltages of the former are the conjugates of the corresponding quantities in the latter. For this reason the circuit of Figure 1C at present may seem superfluous. Actually, however, both these circuits are needed simultaneously for the determination of the torques as will be shown in a following section.

EQUIVALENT CIRCUIT FOR ROTOR EXCITATION ONLY

The equivalent circuit from which the steady-state currents and voltages due to rotor excitation alone may be obtained is derived as follows. Let the subscript O be used to designate all quantities due to the rotor excitation.

If the machine is connected to a balanced unexcited system, the line currents will be balanced, but their frequency will correspond to the machine speed. Assuming operation at speed v, the phase currents are as given by equations 15*a*–15*c*; the angular position is as given by equation 16; and the axis currents are as given by equations 17*a* and 17*b*. These latter currents were obtained by means of equations 3 and 3*a* of reference 4.

$$i_a = i_O \cos vt \qquad (15a)$$

$$i_b = i_O \cos (vt-120) \qquad (15b)$$

$$i_c = i_O \cos (vt+120) \qquad (15c)$$

$$\theta = vt-\theta_0 \qquad (16)$$

$$i_{dO} = i_O \cos \theta_0 \qquad (17a)$$

$$i_{qO} = i_O \sin \theta_0 \qquad (17b)$$

Assuming that the d-c rotor excitations in the direct- and quadrature-axis field circuits are I_{fdO} and I_{fqO}, respectively, the stator axis linkages are given by equations 18*a* and 18*b*. These expressions are taken directly from equation 16*a* of reference 11. It is evident that no currents are flowing in any of the additional rotor circuits because the machine now is assumed to be connected to a balanced unexcited system.

$$\psi_{dO} = x_{afd}I_{fdO} - x_d i_{dO} \qquad (18a)$$

$$\psi_{qO} = x_{afq}I_{fqO} - x_q i_{qO} \qquad (18b)$$

Substituting spin relations of the form of equations 2*a* and 2*b* into equations 18*a* and 18*b* gives the spin linkage equations 19*a* and 19*b* with the spin rotor excitations and spin reactances defined by equations 20*a* and 20*b* and equations 21*a* and 21*b*, respectively.

$$\hat{\psi}_{FO} = \hat{I}_{FO} - \hat{x}_F(0)\hat{\imath}_{FO} - \hat{x}_B(0)\hat{\imath}_{BO} \qquad (19a)$$

$$\hat{\psi}_{BO} = \hat{I}_{BO} - \hat{x}_B(0)\hat{\imath}_{FO} - \hat{x}_F(0)\hat{\imath}_{BO} \qquad (19b)$$

$$\hat{I}_{FO} = 0.707(x_{afd}I_{fdO} + jx_{afq}I_{fqO}) \qquad (20a)$$

$$\hat{I}_{BO} = 0.707(x_{afd}I_{fdO} - jx_{afq}I_{fqO}) \qquad (20b)$$

Substituting equations 19*a* and 19*b* back into the fundamental spin relations, equations 6*a* and 6*b*, gives equations 22*a* and 22*b*.

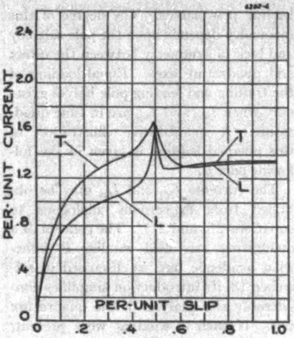

Figure 6. Current in damper bar 1 of machine with incomplete end rings

T—Trailing
L—Leading

$$\hat{x}_F(0) = \frac{x_d(0)+x_q(0)}{2} = \frac{x_d+x_q}{2} \qquad (21a)$$

$$\hat{x}_B(0) = \frac{x_d(0)-x_q(0)}{2} = \frac{x_d-x_q}{2} \qquad (21b)$$

$$\frac{\hat{e}_{FO}}{jv} = \hat{I}_{FO} - \left[\hat{x}_F(0)+\frac{r}{jv}\right]\hat{\imath}_{FO} - \hat{x}_B(0)\hat{\imath}_{BO} \qquad (22a)$$

$$\frac{-\hat{e}_{BO}}{jv} = \hat{I}_{BO} - \hat{x}_B(0)\hat{\imath}_{FO} - \left[\hat{x}_F(0)-\frac{r}{jv}\right]\hat{\imath}_{BO} \qquad (22b)$$

If the machine is connected to a balanced unexcited infinite system, the voltages $\hat{e}_{FO}$ and $\hat{e}_{BO}$ are zero, and the reactance expressions include only the machine reactances. If the system is of finite magnitude, the voltages $\hat{e}_{FO}$ and $\hat{e}_{BO}$ still are to be considered zero, but the system reactance then must be added to the machine reactances. In either case equations 22*a* and 22*b* will reduce to equations 23*a* and 23*b*.

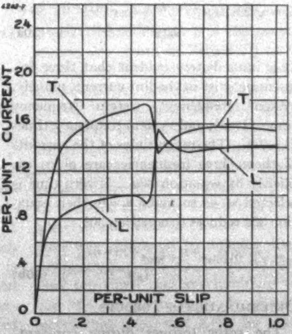

Figure 7. Current in damper bar 2 of machine with incomplete end rings

T—Trailing
L—Leading

$$\hat{I}_{FO} = \left[\hat{x}_F(0)+\frac{r}{jv}\right]\hat{\imath}_{FO} + \hat{x}_B(0)\hat{\imath}_{BO} \qquad (23a)$$

$$\hat{I}_{BO} = \hat{x}_B(0)\hat{\imath}_{FO} + \left[\hat{x}_F(0)-\frac{r}{jv}\right]\hat{\imath}_{BO} \qquad (23b)$$

It is evident that equations 23*a* and 23*b* can be solved electrically by the a-c network of Figure 1A.

DETERMINATION OF ROTOR AND STATOR CURRENTS

The rotor and stator currents during constant-speed asynchronous operation at slip s are obtained from the spin circuits of Figures 1A, 1B, and 1C as follows.

From expressions analogous to equations 2*a* and 2*b*, the spin currents are obtained in terms of the axis currents by means of equations 24*a* and 24*b*.

$$\sqrt{2}i_d = \hat{\imath}_F + \hat{\imath}_B \qquad (24a)$$

$$j\sqrt{2}i_q = \hat{\imath}_F - \hat{\imath}_B \qquad (24b)$$

Substituting equations 12*a* and 12*b* into equations 24*a* and 24*b* and adding the currents due to rotor excitation gives i_d and i_q in terms of the spin-circuit currents as in equations 25*a* and 25*b*.

$$i_{ds} = \frac{\hat{\imath}_{Fs}+\hat{\imath}_{Bs}}{\sqrt{2}}\epsilon^{+jst} + \frac{\hat{\imath}_{Fs}^{*}+\hat{\imath}_{Bs}^{*}}{\sqrt{2}}\epsilon^{-jst} + \frac{\hat{\imath}_{FO}+\hat{\imath}_{BO}}{\sqrt{2}} \qquad (25a)$$

$$i_{qs} = \frac{\hat{\imath}_{Fs}-\hat{\imath}_{Bs}}{j\sqrt{2}}\epsilon^{+jst} - \frac{\hat{\imath}_{Fs}^{*}-\hat{\imath}_{Bs}^{*}}{j\sqrt{2}}\epsilon^{-jst} + \frac{\hat{\imath}_{FO}-\hat{\imath}_{BO}}{j\sqrt{2}} \qquad (25b)$$

The damper-winding currents in the direct and quadrature axes are obtained by means of the above equations and the current moduli $\hat{U}_{nds}$ and $\hat{U}_{nqs}$. These current moduli were introduced in equations 9*a*–9*c* of reference 13. $\hat{U}_{nds}$ is defined as the a-c current (in complex notation) in the nth bar of the direct-axis circuit (Figure 1, reference 13) at slip s when the terminal current i_{ds} of the direct-axis circuit (Figure 1, reference 13) is $1.00\ \underline{/0}$. $\hat{U}_{nqs}$ is defined analogously for the quadrature-axis circuit. The currents in the nth bar of the direct- and quadrature-axis circuits are then as given by equations 26*a*–26*d*. The subzero currents of equations 25*a* and 25*b* obviously cause no currents to flow in the damper windings because the frequency of these currents corresponds to the rotor speed.

$$\hat{U}_{mds} = \hat{I}_{nds} \quad \text{for } \mathfrak{i}_{ds} = 1.00\angle 0 \qquad (26a)$$

$$\hat{U}_{mqs} = \hat{I}_{nqs} \quad \text{for } \mathfrak{i}_{qs} = 1.00\angle 0 \qquad (26b)$$

$$I_{nds} = \frac{\mathfrak{i}_{Fs} + \mathfrak{i}_{Bs}}{\sqrt{2}} \hat{U}_{mds} \epsilon^{+jst} + \frac{\mathfrak{i}_{Fs}^* + \mathfrak{i}_{Bs}^*}{\sqrt{2}} \hat{U}_{nds}^* \epsilon^{-jst} \qquad (26c)$$

$$I_{nqs} = \frac{\mathfrak{i}_{Fs} - \mathfrak{i}_{Bs}}{j\sqrt{2}} \hat{U}_{nqs} \epsilon^{+jst} - \frac{\mathfrak{i}_{Fs}^* - \mathfrak{i}_{Bs}^*}{j\sqrt{2}} \hat{U}_{nqs}^* \epsilon^{-jst} \qquad (26d)$$

I_{nds} and I_{nqs} may be expressed more concisely as in equations 27*a* and 27*b*.

$$I_{nds} = \sqrt{2}\mathfrak{R}[(\mathfrak{i}_{Fs} + \mathfrak{i}_{Bs}) \hat{U}_{nds} \epsilon^{+jst}] \qquad (27a)$$

$$\sqrt{2}\mathfrak{I}[(\mathfrak{i}_{Fs} - \mathfrak{i}_{Bs}) \hat{U}_{nqs} \epsilon^{+jst}] \qquad (27b)$$

Equations 27*a* and 27*b* give the axis bar currents as real functions of time, and complex quantities are not present.

The preceding expressions have given the axis currents. The true currents in the physical bars are given by the following expressions which are taken from equations 9*d* and 9*e* of reference 13.

Bars on trailing pole halves

$$I_{ns} = +I_{nds} + I_{nqs} \qquad (28a)$$

Bars on leading pole halves

$$I_{ns} = -I_{nds} + I_{nqs} \qquad (28b)$$

Due consideration must be given to the bar numbering system described in reference 13 by which the direct-axis bars are numbered from the direct axis and the quadrature axis bars are numbered from the quadrature axis.

It is evident from equations 28*a* and 28*b* that, in general, there is unequal loading of the bars in the trailing and leading pole halves. The degree of this inequality depends upon the pole saliency and lack of symmetry between the direct and quadrature axes. Equal loading of the trailing and leading pole halves exists only when I_{nds} and I_{nqs} are in time quadrature. A quantitative illustration of this unequal loading is given in the following part *B*.

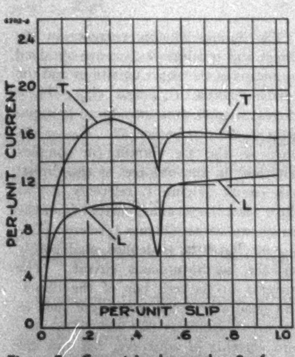

Figure 8. Current in damper bar 3 of machine with incomplete end rings

T—Trailing
L—Leading

The currents I_{fds} and I_{fqs} may be obtained from expressions analogous to equations 27*a* and 27*b*. The concept of a quadrature axis field winding is somewhat academic but, as discussed in reference 13, its introduction simplifies considerably the analysis of the quadrature axis. If such a winding were present, however, it would not be a part of the direct-axis field winding, and expressions such as equations 28*a* and 28*b* could not be written for the field circuit. The alternating currents in the direct and quadrature axis field-winding circuits are given by equations 27*a* and 27*b* directly. The rotor excitation currents I_{fdO} and I_{fqO} must be added to these, of course, to obtain the total current.

The stator line currents are obtained, as follows, directly from the spin currents of Figures 1A and 1B. At a constant slip *s*, the line current i_a in the *a*-phase is expressed in terms of the axis currents as in equations 29*a* and 29*b*. (See equation 14, reference 4.)

$$i_a = i_{ds} \cos \theta - i_{qs} \sin \theta \qquad (29a)$$

$$i_a = i_{ds} \cos [(1-s)t - \theta_0] - i_{qs} \sin [(1-s)t - \theta_0] \qquad (29b)$$

Substituting equations 25*a* and 25*b* into equations 29*a* and 29*b* and noting that $\mathfrak{i}_{FO} = \mathfrak{i}_{BO}^*$ simplifies the latter into equation 30*a*.

$$i_a = \sqrt{2}\mathfrak{R}[\mathfrak{i}_{Fs} \epsilon^{+(jt-\theta_0)} + \mathfrak{i}_{FO} \epsilon^{+j[(1-s)t-\theta_0]} + \mathfrak{i}_{Bs} \epsilon^{-j[(1-2s)t-\theta_0]}] \qquad (30a)$$

It is immediately evident that three frequencies exist in the line current, namely, system frequency, system frequency times $(1-s)$, and system frequency times $(1-2s)$. The magnitudes of the currents of these three frequencies are given explicitly by equation 30*a*. If zero time is selected so as to make $\theta_0 = 0$, then equation 30*a* reduces to equation 30*b*.

$$i_a = \sqrt{2}\ \mathfrak{R}[\mathfrak{i}_{Fs} \epsilon^{+jt} + \mathfrak{i}_{FO} \epsilon^{+j(1-s)t} + \mathfrak{i}_{Bs} \epsilon^{-j(1-2s)t}] \qquad (30b)$$

DETERMINATION OF TORQUES

The rotor torques during constant-speed asynchronous operation at slip *s* are obtained from the spin circuits of Figure 1 as follows.

In terms of the axis quantities, the rotor torque is given by equation 31*a*. (See equation 19, reference 4.)

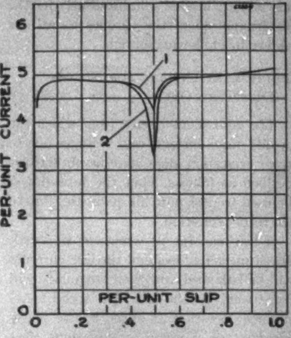

Figure 9. Slip-frequency current in direct-axis field winding

1—Complete end rings
2—Incomplete end rings
The exciting current I_{fdo} must be added to these curves to obtain the total current

$$T = i_q \psi_d - i_d \psi_q \qquad (31a)$$

Substituting the spin quantities defined by equations 2*a* and 2*b* and analogous expressions into equation 31*a* will give the general torque equation in terms of spin quantities.

$$T = j[\mathfrak{i}_B \hat{\psi}_F - \mathfrak{i}_F \hat{\psi}_B] \qquad (31b)$$

The currents and linkages used in equation 31*b* must include the effects of both rotor and stator excitation. The linkages due to the stator excitation are given by equations 4*a* and 4*b*; the currents due to the stator excitation are given by equations 12*a* and 12*b*; the linkages due to the rotor excitation are given by equations 19*a* and 19*b*; and the currents due to the rotor excitation are simply $\mathfrak{i}_{FO}$ and $\mathfrak{i}_{BO}$. Substituting these equations into equation 31*b* will give the torque expressions of equations 32*a*–32*g*, 33*a*–33*f*, and 34*a*–34*d*. Note that the torques now are known directly in terms of the currents and voltages of the spin circuits of Figures 1A, 1B, and 1C.

Zero-Frequency Torque T_{0s}

$$-T_{0s} = (V_{00} - V_{0^*0^*}) + (V_{11} - V_{22}) - (V_{1^*1^*} - V_{2^*2^*}) \qquad (32a)$$

$$(\hat{e}_{FO})^* \mathfrak{i}_{FO} = 0 + jV_{00} \qquad (32b)$$

$$(\hat{e}_{BO})^* \mathfrak{i}_{BO} = 0 + jV_{0^*0^*} \qquad (32c)$$

$$(\hat{e}_{Fs})^* \mathfrak{i}_{Fs} = W_{11} + jV_{11} \qquad (32d)$$

$$(\hat{e}_{Fs}^*)^* \mathfrak{i}_{Fs}^* = W_{1^*1^*} + jV_{1^*1^*} \qquad (32e)$$

$$(\hat{e}_{Bs})^* \mathfrak{i}_{Bs} = W_{22} + jV_{22} \qquad (32f)$$

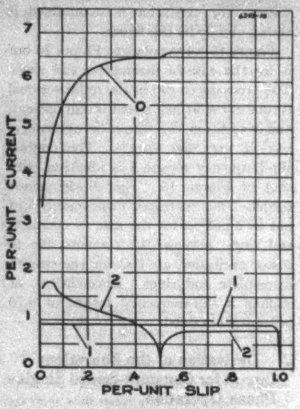

Figure 10. Stator current in machine with complete end rings

0—Current of system frequency
1—Current of system frequency times $(1-s)$
2—Current of system frequency times $(1-2s)$

$$(\hat{e}_{B_s}{}^*)^*i_{B_s}{}^*=W_{2^*2^*}+jV_{2^*2^*} \qquad (32g)$$

Slip-Frequency Torque T_{1s}

$$T_{1s}=2\sqrt{(W_{01}-W_{0^*2}+W_{20^*}-W_{10})^2+(V_{01}-V_{0^*2}-V_{20^*}+V_{10})^2} \qquad (33a)$$

$$-T_{1s}=2(W_{01}-W_{0^*2}+W_{20^*}-W_{10})\times \sin st+2(V_{01}-V_{0^*2}-V_{20^*}+V_{10})\times \cos st \qquad (33b)$$

$$(\hat{e}_{FO})^*i_{F_s}=W_{01}+jV_{01} \qquad (33c)$$

$$(\hat{e}_{BO})^*i_{B_s}=W_{0^*2}+jV_{0^*2} \qquad (33d)$$

$$(\hat{e}_{F_s})^*i_{FO}=W_{10}+jV_{10} \qquad (33e)$$

$$(\hat{e}_{B_s})^*i_{BO}=W_{20^*}+jV_{20^*} \qquad (33f)$$

Double-Slip-Frequency Torque T_{2s}

$$T_{2s}=2\sqrt{(W_{12^*}-W_{21^*})^2+(V_{12^*}-V_{21^*})^2} \qquad (34a)$$

$$T_{2s}=2(W_{12^*}-W_{21^*})\times \sin 2st-2(V_{12^*}-V_{21^*}) \cos 2st \qquad (34b)$$

$$(\hat{e}_{F_s})^*i_{B_s}{}^*=W_{12^*}+jV_{12^*} \qquad (34c)$$

$$(\hat{e}_{B_s})^*i_{F_s}{}^*=W_{21^*}+jV_{21^*} \qquad (34d)$$

The method of writing the products of complex numbers as used above is necessary in order to clarify the $W+jV$ readings taken on the network analyzer. If the analyzer flow conventions are such that positive flow of reactive power is opposite to the positive flow of real power, then the analyzer is wired internally so that the product e^*i is obtained by plugging the voltage e (not its conjugate) to the voltage terminals and the current i to the current terminals. In order to maintain the forementioned flow convention, the $W+jV$ meter will reverse the angle of the voltage automatically and will give e^*i. Hence, on the voltages in the foregoing expressions, the (*) outside the parentheses refers to the conjugation automatically performed by the meter. The potential terminals of the meter are to be plugged to the voltage inside the parentheses.

The forementioned currents and voltages are indicated on the spin circuits of Figures 1A, 1B, and 1C. The determination of the various torques during constant-speed-asynchronous operation has been reduced to the measurement of the real and reactive power in these spin circuits.

B. Numerical Example of the Application of the Equivalent Circuits of Part A

As a numerical example illustrating the application of the equivalent circuits developed in part *A*, part *B* will present the calculated operating characteristics of the synchronous machines described in the following paragraph when operating asynchronously on an infinite system.

SYNCHRONOUS MACHINE

937.5 kva, 750 kw
779 volts line-line
695 amperes line current
6-pole, 1,200-rpm
3-phase, 60-cycle
six damper bars per pole

$e_{ao}=636$ volts per phase, peak
$i_{ao}=983$ amperes line, peak
$x_{ao}=0.647$ ohm

$r=0.008$
$x_d=1.093$
$x_q=0.587$

The impedance functions $\mathfrak{x}_d(js)$ and $\mathfrak{x}_q(js)$, as defined by equation 8*b* of reference 13 and Figure 1 of the same reference, are shown in Figure 2B as functions of the per-unit slip. Two end-ring conditions are shown by these curves. Curves 1 and 2 show $\mathfrak{x}_d(js)$ and $\mathfrak{x}_q(js)$, respectively, for the case of complete end rings, and curves 1 and 3 show the same functions for the case of incomplete end rings. Since it is physically impossible to open the end rings between poles completely, the incomplete end ring condition was simulated by making the impedances X_{e11_q} and R_{e11_q} of the incomplete ring ten times the corresponding values for the complete ring. It should be noted that $\dot{\mathfrak{x}}_d(js)$ and $\dot{\mathfrak{x}}_q(js)$ for unit slip are in satisfactory agreement with normal subtransient reactance values.

The impedance operators $\mathfrak{x}_d(js)$ and $\mathfrak{x}_q(js)$ as given in Figure 2B were obtained by a-c network analyzer measurements of the equivalent circuits shown in Figure 1 of reference 13. The impedances used in these circuits were calculated by the per-unit impedance formulas of the same reference. These per-unit formulas, in turn, are based upon the x_{ad} base-current ratio recommended in reference 12 as a preferred standard, and evaluated in equations 1*c* and 1*d* of the same reference. These base-current ratios are as follows.

$$\frac{I_{fdo}}{(^3/_2)i_{ao}}=\frac{4}{\pi}\frac{A_{d1}}{F_{d1}}\frac{N}{K_pK_dP_1}\frac{1}{N_{fd}} \qquad (35a)$$

$$\frac{I_{zdo}}{(^3/_2)i_{ao}}=\frac{4}{\pi}\frac{A_{d1}}{D_{d1z}}\frac{N}{K_pK_dP_1}\frac{1}{N_{zd}} \qquad (35b)$$

Substituting the required design constants in the foregoing formulas gives the following for the unit currents in the field and damper windings:

$$I_{fdo}=24.70 \text{ amperes} \qquad (35c)$$

$$I_{zdo}=5{,}300 \text{ amperes} \qquad (35d)$$

As discussed in reference 13, I_{zdo} is the base current for the damper-winding circuits in both the direct and quadrature axes. By this means, the per-unit values of I_d and I_q may be added directly.

The operating characteristics of the two synchronous machines defined by the foregoing constants and the impedance operators of Figure 2 are given by Figures 3–15 inclusive. These characteristics describe asynchronous operation at a constant slip, with the direct-axis field winding carrying unit excitation, with no quadrature axis field winding, and with

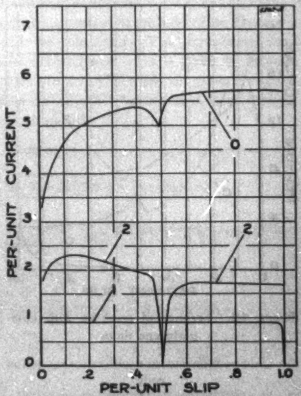

Figure 11. Stator current in machine with incomplete end rings

0—Current of system frequency
1—Current of system frequency times $(1-s)$
2—Current of system frequency times $(1-2s)$

the machines connected to an infinite system of unit voltage and synchronous frequency.

Figures 3, 4, and 5 show the magnitudes of the slip-frequency currents in damper bars 1, 2, and 3, respectively, for the machine with complete end rings; the curves marked *T* and *L* refer to the trailing and leading bars, respectively. The numbering system which defines the physical location of the damper bars in the pole is given on Figure 2A. It should be noted from Figures 3, 4, and 5 that above half speed, the trailing bar is loaded more heavily while below half speed the leading bar carries a heavier load. The inequality is greater above half speed, however, and the trailing bar in general must carry a much heavier load than the leading bar.

Figures 6, 7, and 8 show corresponding curves for the machine with incomplete end rings. In this case, the trailing bar carries a heavier load than the leading bar almost continuously, this inequality being especially marked for bar 2 and bar 3. The unequal heating which is concomitant with this unequal current distribution influences the design of the damper winding to an appreciable degree.

Figure 9 shows the slip-frequency current in the direct-axis field windings for both complete and incomplete end rings. Except at the half-speed point, there is little difference between the two cases. It should be remembered that to obtain the total current in the field winding, the rotor-excitation current $I_{fd0}=1.0$ must be added to the curves of Figure 9.

Figures 10 and 11 show the stator-current curves for complete and incomplete end rings, respectively. The system-frequency current for the case of incomplete end rings is less than the same current for the case of complete end rings almost in the ratio of the average of the $_{xd}(j\omega)$ and $_{xq}(j\omega)$ impedances, as expected. The currents labeled 2, which are of system frequency times $(1-2s)$, are larger for the case of incomplete end rings due to the greater inequality between the *d* and *q* axes. The currents labeled 1, which are of system frequency times $(1-s)$, are exactly the same for the two cases because these currents are caused by the rotor excitation only.

The torque characteristics are given by Figures 12–15, inclusive. Figure 12 shows the torques of the machine with complete end rings with stator and rotor excitation applied separately, but not concurrently. The curve marked 0 gives the unidirectional torque caused by stator excitation alone, and the curve marked 0′ gives the unidirectional torque due to rotor excitation alone. These two torques are naturally opposite in sign. The small dip in the 0 curve around half speed indicates little tendency to lock at this speed. The curve marked 2 gives the double-slip-frequency torque due to stator excitation alone. Note that this torque exceeds the maximum of the unidirectional torque.

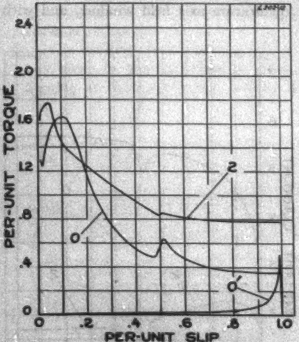

Figure 12. Torques of machine with complete end rings

0—Unidirectional torque with stator excitation only

0—Unidirectional torque with rotor excitation only

2—Double-slip-frequency torque with stator excitation only

The unidirectional torques 0 and 0′ are opposite in sign

Figure 13 corresponds to Figure 12 but is for the case of incomplete end rings. Note that the unidirectional torque 0 due to stator excitation alone is less than that of Figure 12 and has an extreme dip at half speed. The double-slip-frequency torque 2 is very large, indicating extreme torque pulsations. The unidirectional torque 0′ due to rotor excitation alone is identical with the corresponding curve of Figure 12.

Figure 14 shows the effect of applying rotor and stator excitation concurrently for the machine with complete end rings. The unidirectional torque 0 is similar to that of Figure 12 but becomes negative at very low speeds. The double-slip-frequency torque 2 is identical with that of Figure 12. The slip-frequency torque 1, however, has increased considerably, because the flux caused by rotor excitation has reacted with the currents resulting from stator excitation. At standstill this slip-frequency torque reaches a value of 5.8 per unit. The danger in inadvertently stalling a synchronous machine while rotor excitation is maintained is shown vividly by this figure.

Figure 15 corresponds to Figure 14 but is for the case of incomplete end rings. The pronounced dip in the unidirectional torque curve at half speed is still present, and in addition, this torque becomes negative at very low speeds. The double-slip-frequency torque 2 is identical with that of Figure 13. The slip-frequency torque 1 has increased considerably but not as much as for the case of complete end rings, because the currents due to stator excitation are less for the case of incomplete end rings. Nevertheless, at stalling a slip-frequency torque of 4.70 per unit is obtained.

C. Derivation of the Equivalent Circuit for Constant-Speed Single-Phase Operation

Derivation of Equivalent Circuit

The equivalent circuit from which the machine characteristics during constant-speed single-phase operation can be obtained will be derived in this section. The single-phase operation studied in this section is of the type shown in Figure 16. A load is connected across two of the machine terminals, but the machine is not connected to any system. The single-phase load is simulated by putting an impedance *Z* in each line and then tying two phases together with the third open. The

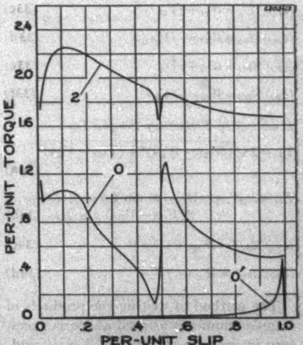

Figure 13. Torques of machine with incomplete end rings

0—Unidirectional torque with stator excitation only

0′—Unidirectional torque with rotor excitation only

2—Double-slip-frequency torque with stator excitation only

The unidirectional torques 0 and 0′ are opposite in sign

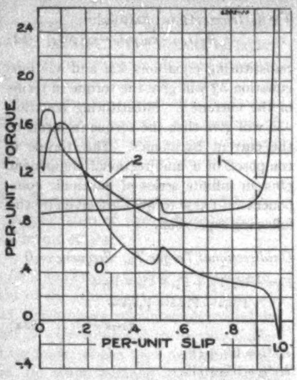

Figure 14. Torques of machine with complete end rings with stator and rotor excitations of 1.0 each applied concurrently

0—Unidirectional torque
1—Slip-frequency torque
2—Double-slip-frequency torque
The slip-frequency torque reaches a maximum of 5.80 at 1.00 slip

load impedance Z is lumped with the machine impedance, and the machine terminal voltage discussed below is the voltage external to the load impedance. Such single-phase operation is defined by equations 36*a*–36*d*.

$$i_a = 0 \tag{36a}$$

$$i_b = -i_c = i \tag{36b}$$

$$e_a = e_a \tag{36c}$$

$$e_b = e_c = e \tag{36d}$$

The axis linkages are given by equations 37*a* and 37*b*. Note that these linkages differ from those of equations 18*a* and 18*b* because the constant rotor excitation in this case can produce currents of frequencies different from that corresponding to the machine speed.

$$\psi_d = x_{afd} I_{fdo} - x_d(p) i_d \tag{37a}$$

$$\psi_q = x_{afq} I_{fqo} - x_q(p) i_q \tag{37b}$$

The spin linkages, as obtained from equations 2*a* and 2*b* and analogous equations, are given by equations 38*a* and 38*b*.

$$\hat{\psi}_F = \hat{I}_{FO} - x_F(p)\hat{i}_F - x_B(p)\hat{i}_B \tag{38a}$$

$$\hat{\psi}_B = \hat{I}_{BO} - x_B(p)\hat{i}_F - x_F(p)\hat{i}_B \tag{38b}$$

Substituting equations 38*a* and 38*b* into the general voltage equations in the spin system, equations 6*a* and 6*b*, gives equations 39*a* and 39*b*.

$$\hat{e}_F = (p + jv)[\hat{I}_{FO} - x_F(p)\hat{i}_F - x_B(p)\hat{i}_B] - r\hat{i}_F \tag{39a}$$

$$\hat{e}_B = (p - jv)[\hat{I}_{BO} - x_B(p)\hat{i}_F - x_F(p)\hat{i}_B] - r\hat{i}_B \tag{39b}$$

The spin currents may be expressed as follows.[14]

$$\hat{i}_F = \frac{\sqrt{2}}{3}[i_a \epsilon^{-j\theta a} + i_b \epsilon^{-j\theta b} + i_c \epsilon^{-j\theta c}] \tag{40a}$$

$$\hat{i}_B = \frac{\sqrt{2}}{3}[i_a \epsilon^{+j\theta a} + i_b \epsilon^{+j\theta b} + i_c \epsilon^{+j\theta c}] \tag{40b}$$

Analogous equations for the spin voltages can be written.

Assuming operation at a constant rotor speed of v per unit, equations 40*a* and 40*b* can be reduced to equations 41*a* and 41*b*.

$$\hat{i}_F = \frac{\sqrt{2}}{3}\epsilon^{-jvt}(i_a + i_b \epsilon^{+j120°} + i_c \epsilon^{-j120°}) \tag{41a}$$

$$\hat{i}_B = \frac{\sqrt{2}}{3}\epsilon^{+jvt}(i_a + i_b \epsilon^{-j120°} + i_c \epsilon^{+j120°}) \tag{41b}$$

Introducing the single-phase definitions of equations 36*a*–36*d* will reduce equations 41*a* and 41*b* and the analogous voltage equations to equations 42*a*–42*d*.

$$\hat{i}_F = +ji\sqrt{^2/_3}\,\epsilon^{-jvt} \tag{42a}$$

$$\hat{i}_B = -ji\sqrt{^2/_3}\,\epsilon^{+jvt} \tag{42b}$$

$$\hat{i}_B = -\hat{i}_F \epsilon^{+j2vt} \tag{42c}$$

$$\hat{e}_B = +\hat{e}_F \epsilon^{+j2vt} \tag{42d}$$

Substituting equations 42*a* and 42*b* into equations 39*a* and 39*b* and equating the two equations by means of equations 42*c* and 42*d* will give the single-phase spin equation 43.

$$\begin{aligned}\epsilon^{+jvt}(p+jv)x_F(p)i\epsilon^{-jvt} - \epsilon^{+jvt}(p+jv)x_B(p) \times \\ i\epsilon^{+jvt} - \epsilon^{-jvt}(p-jv)x_B(p)i\epsilon^{-jvt} + \\ \epsilon^{-jvt}(p-jv)x_F(p)i\epsilon^{+jvt} + 2ri = \\ \sqrt{^3/_2}\,v[\hat{I}_{FO}\epsilon^{+jvt} + \hat{I}_{BO}\epsilon^{-jvt}]\end{aligned} \tag{43}$$

To obtain a solution of equation 43, assume that i can be expressed as in equation 44.

$$\begin{aligned}i = A_1 \cos vt + A_3 \cos 3vt + \ldots + \\ A_m \cos mvt + \ldots + A_1' \sin vt + \\ A_3' \sin 3vt + \ldots + A_m' \sin mvt + \ldots\end{aligned} \tag{44}$$

The assumed current i can be expressed in complex form as in equation 45*a*, with the complex coefficients defined as in equations 45*b* and 45*c* in terms of the real coefficients of equation 44.

$$\begin{aligned}i = \hat{i}_1\epsilon^{+jvt} + \hat{i}_3\epsilon^{+j3vt} + \ldots + \hat{i}_m\epsilon^{+jmvt} + \ldots + \\ \hat{i}_1{}^*\epsilon^{-jvt} + \hat{i}_3{}^*\epsilon^{-j3vt} + \ldots + \hat{i}_m{}^*\epsilon^{-jmvt} \\ + \ldots\end{aligned} \tag{45a}$$

$$\hat{i}_m = \frac{A_m - jA_m'}{2} \tag{45b}$$

$$\hat{i}_m{}^* = \frac{A_m + jA_m'}{2} \tag{45c}$$

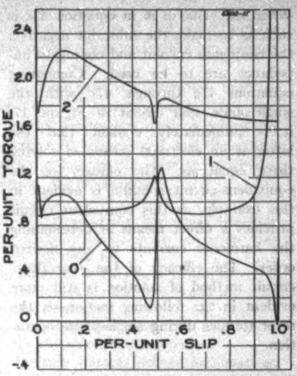

Figure 15. Torques of machine with incomplete end rings with stator and rotor excitations of 1.0 each applied concurrently

0—Unidirectional torque
1—Slip-frequency torque
2—Double-slip-frequency torque
The slip-frequency torque reaches a maximum of 4.70 at 1.00 slip

Consider the general term given by equation 46. Note that for negative values of m, the conjugate coefficient $i_m{}^*$ is to be used instead of i_m.

$$i = \hat{i}_m \epsilon^{+jmvt} \tag{46}$$

Substitute equation 46 into the single-phase spin equation 43, group together the terms containing like exponentials, and equate these groups to zero. An infinite number of equations, corresponding to the infinite number of values of m thereby will be obtained. The equations for $m = +1$ and $m = -1$ are given by equations 47*a* and 47*b*, while the general equation for all other values of m is given by equation 47*c*.

Let $m = +1$

$$\begin{aligned}\left[\mathfrak{x}_F(0) + \mathfrak{x}_F(j2v) + \frac{2r}{jv}\right]\hat{i}_1 - \mathfrak{x}_B(0)\hat{i}_1{}^* - \\ \mathfrak{x}_B(j2v)\hat{i}_3 = \frac{1}{j}\sqrt{\frac{3}{2}}\hat{I}_{FO}\end{aligned} \tag{47a}$$

Let $m = -1$

$$\begin{aligned}\left[\mathfrak{x}_F(0) + \mathfrak{x}_F{}^*(j2v) - \frac{2r}{jv}\right]\hat{i}_1{}^* - \mathfrak{x}_B(0)\hat{i}_1 - \\ \mathfrak{x}_B{}^*(j2v)\hat{i}_3{}^* = j\sqrt{\frac{3}{2}}\hat{I}_{BO}\end{aligned} \tag{47b}$$

Let $m = m$ (positive or negative)

$$\begin{aligned}\left\{\mathfrak{x}_F[j(m-1)v] + \mathfrak{x}_F[j(m+1)v] + \frac{2r}{jmv}\right\}\hat{i}_m - \\ \mathfrak{x}_B[j(m-1)v]\hat{i}_{m-2} - \mathfrak{x}_B[j(m+1)v] \times \\ \hat{i}_{m+2} = 0\end{aligned} \tag{47c}$$

When the value of m in equation 47*c* is such as to make the subscript negative, the conjugate current and conjugate impedance are to be used. Comparing equations 47*a* through 47*c* with the single-phase spin circuit of Figure 17, it is immediately evident that the latter is an electrical means of solving these equations. The efficacy of the equivalent-circuit solution is evident in this example, as this equivalent circuit provides a direct means of determining the harmonic currents to any desired order. The efficacy of the equivalent-circuit method of solution is still more evident in the following section on the rotor torques during single-phase operation.

DETERMINATION OF ROTOR AND STATOR CURRENTS

Assume that the following currents and their conjugates have been obtained from the single-phasing circuit of Figure 17.

$$\mathfrak{i}_1 = i_1\underline{/\delta_1} \qquad (48a)$$

$$\mathfrak{i}_3 = i_3\underline{/\delta_3} \qquad (48b)$$

$$\mathfrak{i}_m = i_m\underline{/\delta_m} \qquad (48c)$$

Substituting equations 48*a*–48*c* into equation 45*a* will give the stator current as in equations 49*a* and 49*b*.

$$i = 2\Re[\mathfrak{i}_1\epsilon^{+jvt} + \mathfrak{i}_3\epsilon^{+j3vt} + \ldots + \mathfrak{i}_m\epsilon^{+jmvt} + \ldots] \qquad (49a)$$

$$i = 2i_1\cos(vt+\delta_1) + 2i_3\cos(3vt+\delta_3) + \ldots \qquad (49b)$$

The axis currents in the damper bars are obtained by means of i_d, i_q, $\hat{U}_{ndm}$, and $\hat{U}_{nqm}$ as follows. Substituting equations 48*a*–48*c*, 42*a* and 42*b* back into the current equations analogous to equations 2*a* and 2*b* will give the axis currents equations 50*a* and 50*b*.

$$i_d = \frac{1}{j\sqrt{3}}[-\mathfrak{i}_1 + (\mathfrak{i}_1-\mathfrak{i}_3)\epsilon^{+j2vt} + (\mathfrak{i}_3-\mathfrak{i}_5)\epsilon^{+j4vt} + \ldots + \mathfrak{i}_1^* - (\mathfrak{i}_1^*-\mathfrak{i}_3^*)\epsilon^{-j2vt} - (\mathfrak{i}_3^*-\mathfrak{i}_5^*)\epsilon^{-j4vt} - \ldots] \qquad (50a)$$

$$i_q = \frac{1}{\sqrt{3}}[+\mathfrak{i}_1 + (\mathfrak{i}_1+\mathfrak{i}_3)\epsilon^{+j2vt} + (\mathfrak{i}_3+\mathfrak{i}_5)\epsilon^{+j4vt} + \ldots + \mathfrak{i}_1^* + (\mathfrak{i}_1^*+\mathfrak{i}_3^*)\epsilon^{-j2vt} + (\mathfrak{i}_3^*+\mathfrak{i}_5^*)\epsilon^{-j4vt} + \ldots] \qquad (50b)$$

Introducing the current moduli $\hat{U}_{ndm}$ and $\hat{U}_{nqm}$ permits the axis currents in the damper bars to be written as in equations 51*a* and 51*b*.

$$I_{nd} = \frac{2}{\sqrt{3}}\mathfrak{J}[(\mathfrak{i}_1-\mathfrak{i}_3)\hat{U}_{nd2}\epsilon^{+j2vt} + (\mathfrak{i}_3-\mathfrak{i}_5)\hat{U}_{nd4}\epsilon^{+j4vt} + (\mathfrak{i}_5-\mathfrak{i}_7)\hat{U}_{nd6}\epsilon^{+j6vt} + \ldots] \qquad (51a)$$

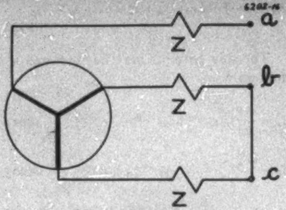

Figure 16. Single-phase loading

Z—System impedance

$$I_{nq} = \frac{2}{\sqrt{3}}\Re[(\mathfrak{i}_1+\mathfrak{i}_3)\hat{U}_{nq2}\epsilon^{+j2vt} + (\mathfrak{i}_3+\mathfrak{i}_5)\hat{U}_{nq4}\epsilon^{+j4vt} + (\mathfrak{i}_5+\mathfrak{i}_7)\hat{U}_{nq6}\epsilon^{+j6vt} + \ldots] \qquad (51b)$$

The physical currents in the various damper bars are obtained from the axis currents equations 51*a* and 51*b* by means of equations 28*a* and 28*b*. The subscript *s*, denoting slip operation, may be dropped from these equations when using equations 51*a* and 51*b*.

DETERMINATION OF TORQUES

The rotor torques during constant-speed single-phase operation can be obtained from the spin circuit of Figure 17.

Substituting the spin linkages of equations 38*a* and 38*b* into the general spin torque expression, equation 31*b*, will give the single-phase torque expression, equation 52.

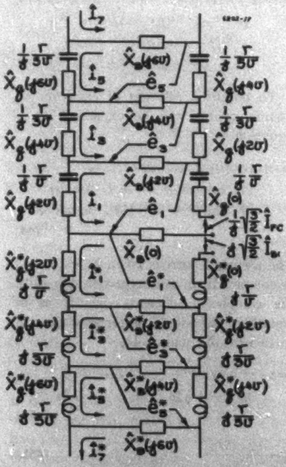

Figure 17. Equivalent circuit for constant-speed single-phase operation of the type defined by Figure 16

$$T = j\mathfrak{i}_B[\hat{I}_{FO} - x_F(p)\mathfrak{i}_F - x_B(p)\mathfrak{i}_B] - j\mathfrak{i}_F[\hat{I}_{BO} - x_B(p)\mathfrak{i}_F - x_F(p)\mathfrak{i}_B] \qquad (52)$$

Substituting equations 42*a* and 42*b* into equation 52 will give the torque in terms of the current i. Substituting equation 45*a* will give then the torque in terms of the current harmonics. This torque is composed of a unidirectional component plus an infinite series of harmonic components. These torques are given by the following expression.

Unidirectional Torque T_0; Frequency = 0

$$T_0 = -{}^4/_3[V_{11} + V_{33} + V_{55} + V_{77} + \ldots - V_{1^*1^*} - V_{3^*3^*} - V_{5^*5^*} - V_{7^*7^*} - \ldots] \qquad (53a)$$

$$(\hat{e}_1)^*\mathfrak{i}_1 = W_{11} + jV_{11}$$
$$(\hat{e}_3)^*\mathfrak{i}_3 = W_{33} + jV_{33}$$
$$(\hat{e}_5)^*\mathfrak{i}_5 = W_{55} + jV_{55}$$
$$(\hat{e}_7)^*\mathfrak{i}_7 = W_{77} + jV_{77}$$
$$(\hat{e}_1^*)^*\mathfrak{i}_1^* = W_{1^*1^*} + jV_{1^*1^*}$$
$$(\hat{e}_3^*)^*\mathfrak{i}_3^* = W_{3^*3^*} + jV_{3^*3^*}$$
$$(\hat{e}_5^*)^*\mathfrak{i}_5^* = W_{5^*5^*} + jV_{5^*5^*}$$
$$(\hat{e}_7^*)^*\mathfrak{i}_7^* = W_{7^*7^*} + jV_{7^*7^*}$$

Second-Harmonic Torque T_2; Frequency = 2vt

$$T_2 = {}^8/_3\sqrt{[W_{13} + W_{35} + \ldots - W_{1^*1} - W_{3^*1^*}\ldots]^2 + [V_{13} + V_{35} + \ldots - V_{1^*1} - V_{3^*1^*} - \ldots]^2} \qquad (54a)$$

$$(\hat{e}_1)^*\mathfrak{i}_3 = W_{13} + jV_{13}$$
$$(\hat{e}_3)^*\mathfrak{i}_5 = W_{35} + jV_{35}$$
$$(\hat{e}_5)^*\mathfrak{i}_7 = W_{57} + jV_{57}$$
$$(\hat{e}_7)^*\mathfrak{i}_9 = W_{79} + jV_{77}$$
$$(\hat{e}_1^*)^*\mathfrak{i}_1 = W_{1^*1} + jV_{1^*1}$$
$$(\hat{e}_3^*)^*\mathfrak{i}_1^* = W_{3^*1^*} + jV_{3^*1^*}$$
$$(\hat{e}_5^*)^*\mathfrak{i}_3^* = W_{5^*3^*} + jV_{5^*3^*}$$
$$(\hat{e}_7^*)^*\mathfrak{i}_5^* = W_{7^*5^*} + jV_{7^*5^*}$$

$$T_2 = -{}^8/_3[(W_{13} + W_{35} + \ldots - W_{1^*1} - W_{3^*1^*} - \ldots)\sin 2vt + (V_{13} + V_{35} + \ldots - V_{1^*1} - V_{3^*1^*} - \ldots)\cos 2vt]$$

Fourth-Harmonic Torque T_4; Frequency 4vt

$$T_4 = {}^8/_3\sqrt{[W_{15} + W_{37} + \ldots - W_{1^*3} - W_{3^*1} - \ldots]^2 + [V_{15} + V_{37} + \ldots - V_{1^*3} - V_{3^*1} - \ldots]^2} \qquad (55a)$$

$$(\hat{e}_1)^*\mathfrak{i}_5 = W_{15} + jV_{15}$$
$$(\hat{e}_3)^*\mathfrak{i}_7 = W_{37} + jV_{37}$$
$$(\hat{e}_5)^*\mathfrak{i}_9 = W_{59} + jV_{59}$$
$$(\hat{e}_1^*)^*\mathfrak{i}_3 = W_{1^*3} + jV_{1^*3}$$
$$(\hat{e}_3^*)^*\mathfrak{i}_1 = W_{3^*1} + jV_{3^*1}$$
$$(\hat{e}_5^*)^*\mathfrak{i}_1^* = W_{5^*1^*} + jV_{5^*1^*}$$

$$T_4 = -{}^8/_3[(W_{15} + W_{37} + \ldots - W_{1^*3} - W_{3^*1} - \ldots)\sin 4vt + (V_{15} + V_{37} + \ldots - V_{1^*3} - V_{3^*1} - \ldots)\cos 4vt]$$

The forementioned currents and voltages are indicated on the spin circuit of Figure 17, and the determination of the torques during constant-speed single-phase operation is reduced to the direct measurement of real and reactive power in this network. The torques of higher harmonic order can be obtained from the preceding equations by analogy.

The discussion given after equations 34*c* and 34*d* concerning the method of writing the products of complex numbers also applies to the above expressions.

D. Numerical Example of the Application of the Equivalent Circuit of Part C

As a numerical example of the application of the equivalent circuit of part *C*, the synchronous machines described in part *B* were subjected to various single-phase loads, of the type defined by Figure 16, while running at synchronous speed. The torques and rotor and stator currents so obtained are given below. In each case, the field current was adjusted to give unit fundamental stator current through the single-phase loads. The single-phase loads were of unit magnitude (per phase) of either pure reactance or pure resistance.

COMPLETE END RINGS, PURE REACTANCE LOAD

Rotor d-c excitation $I_{fd0}=1.85$
Stator line current i = 1.00 (60 cycles)
Rotor-circuit currents
bar 1, trailing = 0.196 (120 cycles)
leading = 0.182 (120 cycles)
bar 2, trailing = 0.194 (120 cycles)
leading = 0.174 (120 cycles)
bar 3, trailing = 0.190 (120 cycles)
leading = 0.114 (120 cycles)
field winding = 0.407 (120 cycles)
Second-harmonic torque = 0.758 (120 cycles)

COMPLETE END RINGS, PURE RESISTANCE LOAD

Rotor d-c excitation $I_{fd0}=1.31$
Stator line current i = 1.00 (60 cycles)
Rotor-circuit currents
bar 1, trailing = 0.207 (120 cycles)
leading = 0.194 (120 cycles)
bar 2, trailing = 0.195 (120 cycles)
leading = 0.173 (120 cycles)
bar 3, trailing = 0.157 (120 cycles)
leading = 0.143 (120 cycles)
field winding = 0.413 (120 cycles)
Undirectional torque = 0.678 (0 cycles)
Second-harmonic torque = 0.670 (120 cycles)

INCOMPLETE END RINGS, PURE REACTANCE LOAD

Rotor d-c excitation $I_{fd0}=1.87$
Stator line current i = 1.00 (60 cycles)
Rotor-circuit excitations
bar 1, trailing = 0.184 (120 cycles)
leading = 0.169 (120 cycles)
bar 2, trailing = 0.168 (120 cycles)
leading = 0.158 (120 cycles)
bar 3, trailing = 0.118 (120 cycles)
leading = 0.118 (120 cycles)
field winding = 0.411 (120 cycles)
Second-harmonic torque = 0.812 (120 cycles)

INCOMPLETE END RINGS, PURE RESISTANCE LOAD

Rotor d-c excitation $I_{fd0}=1.34$
Stator line current = 1.00 (60 cycles)
0.063 (180 cycles)
Rotor-circuit currents
bar 1, trailing = 0.180 (120 cycles)
= 0.019 (240 cycles)
leading = 0.160 (120 cycles)
= 0.006 (240 cycles)
bar 2, trailing = 0.175 (120 cycles)
= 0.011 (240 cycles)
leading = 0.142 (120 cycles)
= 0.010 (240 cycles)
bar 3, trailing = 0.125 (120 cycles)
= 0.007 (240 cycles)
leading = 0.119 (120 cycles)
= 0.008 (240 cycles)
field winding = 0.425 (120 cycles)
= 0.024 (240 cycles)
Unidirectional torque = 0.700 (0 cycles)
Second-harmonic torque = 0.690 (120 cycles)
Fourth-harmonic torque = 0.071 (240 cycles)

All harmonics of orders higher than those given are negligible, and their magnitudes are less than five per cent of the values given. In the first and third cases the unidirectional torques are zero because the loads are purely reactive.

Nomenclature

All quantities are in per unit unless in boldface type, in which case they are in the ampere-inch-second system (amperes, volts, and inches). Quantities such as N, P_1, and so forth for which it obviously is unnecessary to distinguish the ampere-inch-second values by a special script, are in standard type only.

All vector quantities are indicated by a circumflex (ˆ) over the symbol. All conjugate quantities are indicated by an asterisk (*). The d-c rotor-excitation currents and in parts *A* and *B* the stator current due to the d-c rotor-excitation currents are indicated by the subscript *O*. The subscript *s* indicates constant-speed asynchronous operation at slip *s*, and the subscript *m* (which takes the values 1, 2, 3. . .) identifies the harmonic quantities obtained during constant-speed single-phase operation. The subscript 0 identifies the unit values of the various currents and voltages.

The *d* and *q* functions, by which the various quantities are resolved into components in line with the direct and quadrature axes, are termed the axis quantities.

The axis quantities are grouped into vector quantities which in this paper are identified by the subscript *F* and *B*.

The *F* and *B* quantities are equivalent to vectors rotating at slip speed, and for this reason these quantities are termed the "spin" quantities.

A double subscript rotation is used for the $W+jV$ measurements. The first subscript locates the mesh in which the voltage is to be measured, and the second subscript locates the mesh in which the current is to be measured.

A_1, A_3, A_5 = current harmonics obtained during single-phase operation (see equation 44)
A_{d1} is defined in references 12 and 13
D_{d1x} is defined in references 12 and 13
e = phase voltage
e_a, e_b, e_c = phase voltages
e_d, e_q = axis voltages of reference 4
$\hat{e}_F$, $\hat{e}_B$ = spin voltages defined by equations analogous to equations 2*a* and 2*b*
e_{a0} = peak value of rated stator phase voltage
F_{d1} = defined in references 12 and 13
i = line current
i_a, i_b, i_c = phase currents
i_d, i_q = axis currents of reference 4
$\hat{i}_F$, $\hat{i}_B$ = spin currents defined by equations analogous to equations 2*a* and 2*b*
$\hat{i}_1$, $\hat{i}_3$, $\hat{i}_5$ = current harmonics obtained during single-phase operation (see equation 45*a* and Figure 17)
i_{a0} = peak value of rated line current in amperes
I_{fd}, I_{fq} = direct- and quadrature-axis field-winding currents
I_n = current in *n*th damper bar
I_{nd}, I_{nq} = axis currents in *n*th damper bar
$\hat{I}_F$, $\hat{I}_B$ = spin currents defined in equations 20*a* and 20*b*
I_{fdO} = unit current in amperes for field-winding circuit
I_{xdO} = unit current in amperes for damper bars
$\mathcal{I}$ = imaginary part of . . .; $\mathcal{I}(a+jb)=b$
$j=\sqrt{-1}$
K_p, K_d = pitch and distribution factors (greater than 1.0)
m = order of harmonic during single-phase operation
N = stator series turns per phase
N_{fd}, N_{xd} = turns per pole of field winding and *x*th additional circuit, respectively, ($N_{xd}=1.0$ for damper-winding circuits)
p = differential operator $\dfrac{d}{dt}$
P_1 = number of poles
r = stator resistance per phase
$\mathcal{R}$ = real part of . . .; $\mathcal{R}(a+jb)=a$
s = slip
t = time in radians per second
T = torque
T_{0s}, T_{1s}, T_{2s} = torques obtained during asynchronous operation; unidirectional, slip frequency, and double slip frequency, respectively
T_0, T_2, T_4 = harmonic torques obtained during single-phase operation; unidirectional, second harmonic, and fourth harmonic, respectively

Characteristics of a 400-Mile 230-Kv Series-Capacitor-Compensated Transmission System

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SERIES capacitors have been recognized for some time as a means of increasing the amount of power transmitted over a long transmission line and as a means of increasing the distance that alternating-current power can be transmitted economically.[1] The basic concept of neutralizing any portion of the series inductive reactance of a long transmission system by means of series capacitors is theoretically sound. The practical application of series capacitors for such service introduces several new and interesting problems of system design and operation, which while capable of solution will require the attention of engineers of both the equipment manufacturers and the operating companies. It is the objective of this paper to outline system characteristics encountered when series-capacitor compensation is used, with a view toward solving the problems arising therefrom. No attempt will be made to reexamine the relative cost of transmission with and without series compensation. In fact, the cost of the compensated system will be influenced to a certain degree by the methods used to solve the problems arising through use of series compensation.

A brief discussion of some of the major factors that must be considered in the application of series capacitors on any system will present the problems involved and suggest methods of solution. Series capacitors are connected into a transmission line in series with the line reactance in order to neutralize or compensate a portion or possibly all of the line reactance. With greater capacitive reactance, it is theoretically possible, but for several reasons not practical to compensate also a part of the reactance of the terminal equipment at each end of the line. It would be desirable from an operation and protective-relaying standpoint to insert the capacitors uniformly along the line in a large number of small units; however, it is not practical from an economic standpoint to do so because of high overvoltages on the capacitors during system faults and because of the higher cost of a large number of small capacitor stations.

From the standpoint of electrical performance, assuming capacitors are distributed uniformly along the line and are in operation at all times, a 400-mile line with 50 per cent compensation would become the equivalent of a 200-mile line but would have a higher resistance and more line-charging current than the standard conductor used in the line. The electrical performance of such a 200-mile line could be represented exactly by an equivalent hollow-core conductor of rather large diameter and small cross section which would give both the higher resistance and larger charging current.

System Characteristics During Faults

A simple example consisting of a 400-mile line in which the line reactance is assumed to be 100 per cent compensated, but with no compensation of reactance of terminal equipment, will illustrate some of the compensated system characteristics

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$\hat{U}_{nd}$, $\hat{U}_{nq}$ = current moduli defined in equations 26*a* and 26*b* and in reference 13

v = speed ($v = 1 - s$)

V = reactive-power component of $W + jV$

W = real-power component of $W + jV$

x as a subscript denotes the 100-per-cent-pitch damper-winding circuit; defined in reference 13

x_{afd}, x_{afq} = mutual reactances between stator and field windings in direct and quadrature axes; evaluated in reference 13

x_d, x_q = synchronous reactances; evaluated in reference 13

$x_d(p)$, $x_q(p)$ = operational impedances in the axis system; evaluated in reference 13

$x_F(p)$, $x_B(p)$ = operational impedances in the spin system; defined in equations 5*a* and 5*b*

x_{ao} = base stator impedance in ohms

$x_{ao} = \frac{e_{ao}}{i_{ao}}$

Z = system impedance defined by Figure 16

δ_1, δ_2, δ_3 = phase angles of the harmonic currents obtained during single-phase operation; see equations 48*a*–48*c* and Figure 17

θ = space angle between stator and rotor axes; defined in reference 4

ψ_d, ψ_q = axis linkages defined in reference 4

$\hat{\psi}_F$, $\hat{\psi}_B$ = spin linkages defined by equations 2*a* and 2*b*

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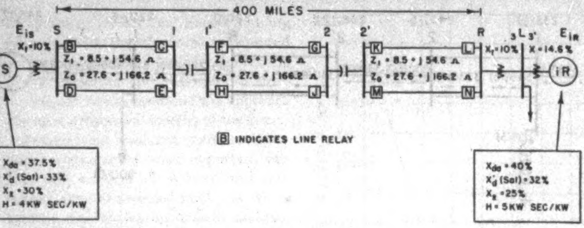

Figure 1. Characteristics of a 400-mile 230-kv compensated transmission system

General conditions

1. Line characteristics
   230 kv, double circuit, 400 miles with 795,000-circular-mil aluminum-cable-steel-reinforced at 27 feet flat spacing and two intermediate sectionalizing stations
2. Terminal equipment rating
   Transformer *R* capacity = kva delivered
   Generator and transformer *S* capacity = rating transformer *R* plus line losses
   Receiver end transformer and generator capacity = 2 times rating transformer *R* (impedances on same base as transformer *R* are $^1/_2$ of those specified)
   Load = approximately 3 times rating transformer *R* at unity power factor
3. Series capacitors
   X_c of compensating capacitors was changed for the various studies but is expressed as a percentage compensation of the total line reactance between buses *S* and *R*, normally with all line sections in service or during emergencies when a faulted line section is out of service
4. Equipment reactances
   X_{da} = adjusted synchronous reactance. Sending end generator is assumed to be hydro-electric. Receiving end generator is assumed to consist of system containing hydro-electric and turbo generation. Reactance of receiver end system (X = 14.6 per cent) assumed to include both transformer and distribution system

Equipment Rating

Case No.	Generator and Transformer *S*	Transformer *R*	Receiver End Generator and Transformer
1–8*	375,000 kva	320,000 kva	640,000 kva
9†	185,000 kva	185,000 kva	370,000 kva

* Ratings based on conductor thermal limit (900 amperes) with one line section out of service.
† Ratings based on static stability limit with 15 per cent margin, with one line section out of service.

during faults. Such a compensated line would have a net line impedance consisting of resistance only, so that any fault current would be limited almost entirely by the impedance of the terminal equipment. Therefore, approximately the same magnitude of fault current would be supplied from the sending-end generators whether a fault was at the generating station high-tension bus or at the receiver bus 400 miles away. In practice full 100-per-cent line compensation probably would not be used; however, this simple example illustrates the fact that fault current tends to be limited more by impedance of terminal equipment rather than by line impedance as the per cent compensation is increased. Hence, if a 100-per-cent-compensated line is connected to a high capacity system having low terminal impedance which could supply a fault current ten times the maximum full load current of the compensated line, then all series capacitors in the line would be subjected to ten times normal voltage for a fault occurring near the receiver bus. Even if the above line is connected to generators and step-up transformers having a total of 40 per cent reactance, the fault current flowing through the capacitors when 100-per-cent compensation is used would result in 2.5 times normal full load capacitor voltage and 6.25 times normal capacitor kilovolt-amperes. In many cases such an increase in momentary kilovolt-ampere capacity and the resultant cost of capacitors cannot be economically justified, so that the capacitors will be installed in two or three large units with protective equipment which short-circuits them as soon as the excessive overvoltage occurs during the fault.

Table I. Emergency Ratings of Capacitors

110 per cent of normal voltage continuously, 121 per cent of normal kilovolt-amperes
135 per cent of normal voltage up to 30 minutes 182 per cent of normal kilovolt-amperes
150 per cent of normal voltage up to 5 minutes, 225 per cent of normal kilovolt-amperes
*200 per cent of normal voltage up to $^1/_2$ cycle, 400 per cent of normal kilovolt-amperes

* Protected by spark gap set at twice normal voltage.

When the capacitors are concentrated in a few large units, they will be placed at intermediate stations in the line as indicated in Figure 1 which shows three line sections but only two capacitor stations. When the per-cent compensation is sufficiently high, the capacitive reactance located in a station becomes greater than the inductive reactance of a line section, because there is always one less capacitor station than the number of line sections. Then, for a fault just beyond a capacitor station a net capacitive reactance is obtained between the fault point and the high-voltage bus of the generating station which causes partial neutralization of the reactance of terminal equipment. This produces a larger fault current for a fault located beyond the capacitor station than for a fault located at the generating station high-voltage bus. For the system shown in Figure 1 a line compensation greater than two-thirds of the line reactance will result in partial neutralization of terminal reactance when the fault is located as described previously. For a 75 per cent compensation of the line and the fault located at point 2', practically all of the sending-end transformer reactance is neutralized by the net capacitive reactance obtained out to the fault point. Similar conclusions are obtained for the components of fault current supplied from the receiver end of the line when the fault occurs on the sending-end side of a capacitor.

Capacitors have quite definite maximum voltage ratings, the magnitude depending on the period during which the capacitor is subjected to overvoltages. Table I specifies four emergency ratings in terms of the allowable voltage, the time interval of application, and the corresponding kilovolt-amperes obtained in the capacitor.

With the two conflicting requirements of a maximum allowable voltage across the capacitor and voltages above normal during system faults, it is desirable for purposes of comparison in this study to assume the capacitor equipment constructed in one of two ways which will be designated hereafter as design *A* or *B*.

A. Construct the capacitor to withstand the maximum momentary voltage to which the capacitors will be subjected during faults. Capacitor cost will increase almost directly with the kilovolt-amperes or as the square of the maximum voltage so that for many cases this design cannot be justified from

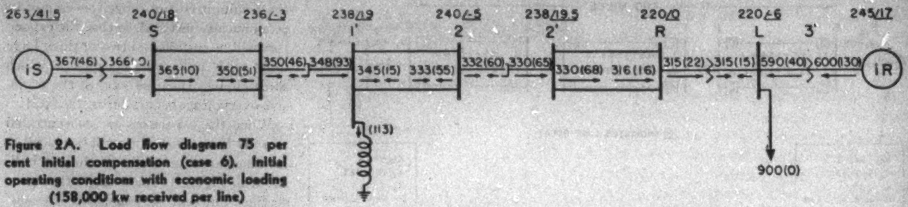

Figure 2A. Load flow diagram 75 per cent initial compensation (case 6). Initial operating conditions with economic loading (158,000 kw received per line)

Power and reactive flow is indicated thus—kw/1,000 (kilovars/1,000)

Arrows in same direction indicate lagging power factor

an economic standpoint. In general, this design should be investigated when the reactance of terminal equipment at each end of the line is relatively high, and when the per-cent compensation is relatively low. When this design is investigated, consideration should be given to the short-time overvoltage ratings of series capacitors. For example, if manufacturers will allow a maximum capacitor voltage of approximately twice rated value during the short period of the fault, a system having a maximum fault current through the capacitors of 2.2 to 2.4 times maximum load current may use this design economically in comparison with the design in which the capacitors are short-circuited during the fault.

B. Provide protective equipment such as a spark gap or short-circuiting breaker to short-circuit the capacitors during a fault, and a contactor or circuit breaker to clear the spark gap as soon as the fault is removed so that the capacitive reactance will be reinserted rapidly in the line. This design would be used when the terminal impedance is low and the per-cent compensation is high. The short-circuiting of the capacitors during the fault plus the effects of the short but important time interval needed to reinsert the capacitors into the line after the fault is cleared make it necessary, in obtaining a given power limit, to use a higher percentage compensation but in general not as high a kilovolt-ampere rating as for design *A* for the same degree of initial line loading.

If design *A* is used, it may be possible and is desirable to have several small capacitor stations in the line. If design *B* is used, the cost of protective equipment dictates the need for a few capacitor stations of relatively large capacity.

Transient Stability Considerations

It can be shown that the most economical loading of a 60-cycle 230-kv 200-mile uncompensated line is approximately 175,000 kw delivered. Similarly, the most economical loading of a 230-kv 400-mile line may be of the order of 165,000 kw delivered. If transient stability is considered, it is possible by proper design to approach the above economical loading on lines of 200-mile length without series compensation, but it is not possible to obtain both stability and the most economical loading on 400-mile lines without using relatively large series compensation. In fact, without compensation on the 400-mile lines, other system design being equal, the transient stability limit may be of the order of 75,000 kw per circuit for a system similar to that shown in Figure 1, except uncompensated. Compensation of lines longer than 400 miles also will show large relative increases in power transmitted without loss of stability.

The determination of the allowable loading of lines and the transient stability limit becomes a major design and operat-

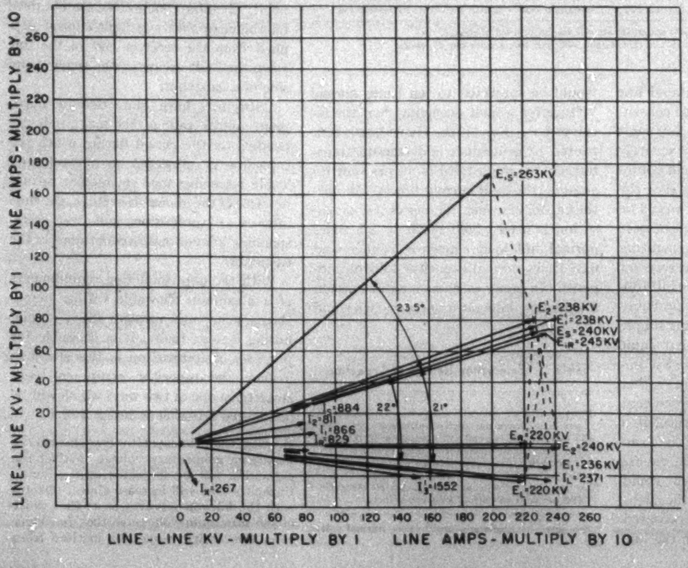

Figure 2B. Vector diagram, 75 per cent initial compensation (case 6). Initial operating conditions with economic loading (158,000 kw received per line)

Equipment rating

Generator *S*—375,000 kva
Transformer *R*—320,000 kva
System *R*—640,000 kva

ing problem for all long lines supplying firm power whether they are compensated or uncompensated. But by increasing the compensation used, the stability-limit loading can be increased and will approach the most economical loading of the line.

Determining transient stability introduces a number of other important factors such as type of system, how connected, type of assumed fault and where located, time of clearing of fault, and percentage of transmission capacity removed when the faulted line section is cleared. Series compensation using design *B* also involves short-circuiting of capacitors during the fault and then reinserting them. If the capacitors are not reinserted at the instant the fault is cleared, a reduction in the stability limit occurs. Assume for example a line with 50-per-cent compensation after the faulty line section has been removed; then for the length of the period of delay in reinserting the capacitors the transmission line reactance is doubled, and this becomes a major factor in maintaining stability. Likewise, less synchronizing power can be transmitted during the period of the fault if the capacitors are out of the circuit than if they remain in the circuit. It is obvious that the period of time the capacitors are short-circuited must be cut to the minimum. The time required to clear the fault, that is, whether sequential tripping must be used or whether it can be assumed that simultaneous tripping can be obtained through a reliable over-all carrier relaying design, is similarly an important factor in determining the stability. To summarize, for design *B* for any specified initial load on the transmission system, there are three important factors which will make it necessary to increase the compensation required to maintain stability, first, removal of the capacitors during the fault; second, increase in fault clearing time; and third, increase in the time required to reinsert the capacitors after the fault is cleared. For design *A* only the second factor applies.

System Analyzed

In order to obtain more specific data on system characteristics resulting when series compensation is used, the system shown in Figure 1 was studied extensively on an a-c calculating board. It will be observed that this system consists of a 2-circuit 230-kv 400-mile line using 795,000-circular-mil aluminum-cable-steel-reinforced conductors with sectionalizing stations at two intermediate points 1–1′ and 2–2′ and with series capacitors placed at these points. It was decided to load the line initially to approximately 900 amperes at the sending end, corresponding to the thermal limit of one line section if the other parallel section were not in service. This assumption also resulted in a delivered power of approximately 160,000 kw per line, which is considered fairly close to the most economical

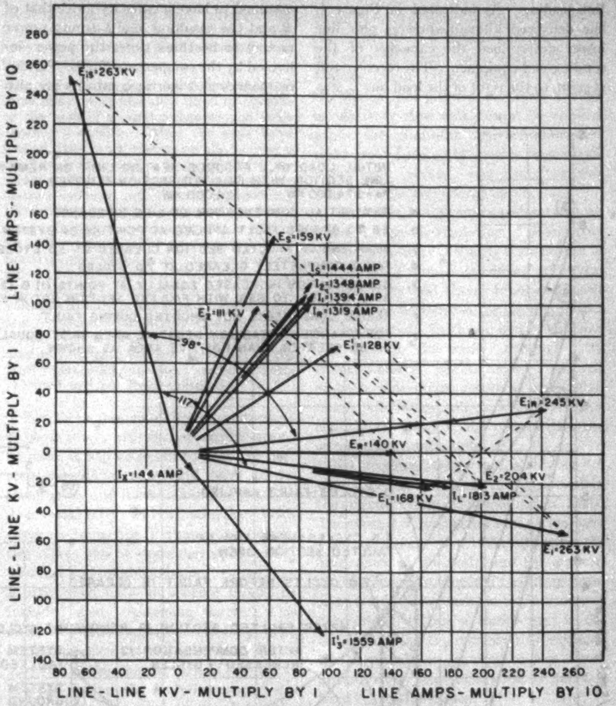

Figure 3. Vector diagram at angle of maximum transient swing (case 6). 75 per cent initial and final compensation

Conditions

1. Initial loading as per Figure 2A
2. 2-phase ground fault on one line near point 1. Near end breaker open at 3.5 cycles. Fault cleared at 7 cycles
3. Capacitors short-circuited during fault periods and reinserted at 10.5 cycles

Angular Difference, Degrees		Ratio $I/I_{initial}$		System Z** Each Line—Ohms		Per Unit Current Through Capacitor	
$E_{iR}-i_S$	98	*S*	3.26*	*S*	110/13.5		
E_{R-S}	66	1	3.22*	1	−188/−62		
$E_{iS}-L$	−113	1′	1.78	1′	178/−21		
$E_{iR}-L$	−15	2	1.66	2	−302/−58	1	1.55†
E_{S-1}	−78	2′	1.66	2′	165/10	2	1.50†
E_S-I_S	−13.5	*R*	1.59	*R*	212/−49		
E_1-I_1	62	*L*	0.77				
$E_1'-I_1'$	21	3′	1.00				
E_2-I_2	58						
$E_2'-I_2$	−10						
E_R-I_R	49						

* Line current carried by single line with faulted section removed.

** Negative *Z* indicates *I* in nontripping direction.
Negative θ indicates capacitive reactance.

† Based on 900-ampere rating.

line loading. As indicated in Figure 1, the delivered kilovolt-amperes plus line losses determined the capacity of the sending-end equipment. The receiver-end system, to the right of the load bus L, was assumed to have a capacity twice that of R and the resultant load L (unity power factor) to be three times the power delivered by the compensated lines. Various reactances and inertia constants are also specified in Figure 1. Load-flow diagrams were obtained with 240-kv sending-end voltage (bus S) and 220-kv load bus voltage (bus L).

Most of the transient stability studies

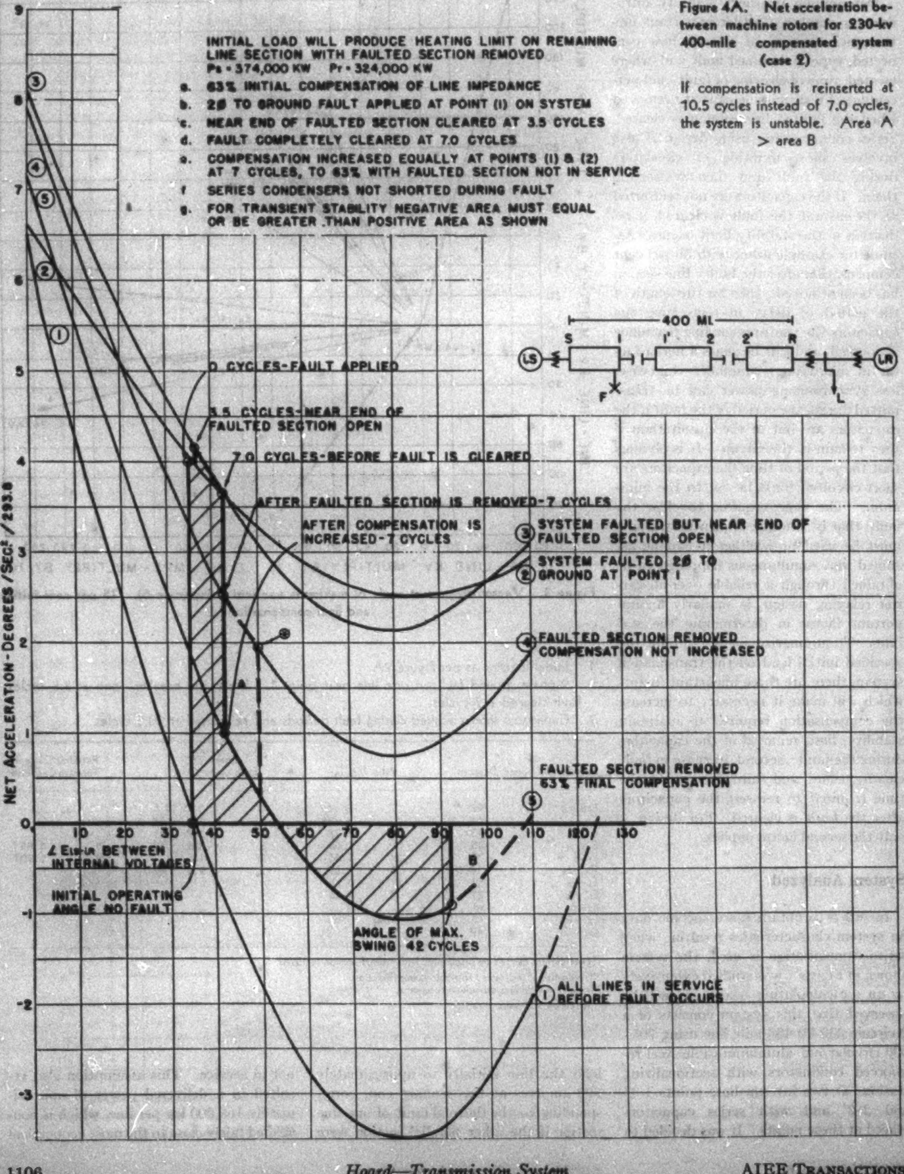

Figure 4A. Net acceleration between machine rotors for 230-kv 400-mile compensated system (case 2)

If compensation is reinserted at 10.5 cycles instead of 7.0 cycles, the system is unstable. Area A > area B assumed a double-line-to-ground fault applied on one line near point 1, Figure 1. Checks were made which showed that the transient shock was more severe for a fault on the sending-end line section as compared to other sections and was slightly more severe for a fault at the sending-end bus than at the point 1 bus. However, because a considerable amount of data had been obtained previously for faults at point 1, it was decided to continue the study with the fault location at this point.

Relaying Characteristics

Both the time required to clear the fault and that required to reinsert the capacitors are of major importance in maintaining stability. These periods are tied up intimately with the protective relaying of the lines and the control of the series capacitors. Of the standard relay schemes available, studies indicate that a high-speed current-balance scheme would provide proper protection for the lines. Other standard relay schemes show limitations which will require major modifications to the schemes before they can be used. In view of the many new factors to be considered, it is planned to make laboratory tests in the future to confirm these results.

It will be shown from examples studied here that relay requirements which must be met, namely, to trip the faulted line section and to block tripping on the unfaulted line sections are such that the standard methods of impedance relaying cannot be used. The following reasons are given:

(*a*). When the compensation is high so that the compensating reactance "reaches" part way into the transformer and generator reactance, faults located beyond the capacitor stations produce greater fault currents, and lower impedances may be seen by impedance relays than when the fault is located near the generator bus.

(*b*). Before the gaps have flashed over, some relays may be called upon to operate on the basis of a small resistance and large capacity reactance. After the gaps have flashed, they must operate on the basis of a small resistance and a large inductive reactance. This produces a shift in phase angle between voltage and current approaching 160 to 170 degrees before and after gap flashover. This system characteristic is critical for a relay scheme using conventional directional relays or their equivalent since it allows only a 10- to 20-degree band between tripping and blocking. Other shifts in phase of voltage and current between a 3-phase fault and unbalanced faults may eliminate completely the simple directional relay since it may not indicate the true direction of the fault when series capacitors are both short-circuited and not short-circuited.

For these reasons if impedance relays or these relays with carrier current are to be used, it will be assumed that the first zone impedance relay will have to be eliminated, and modifications of existing directional relays will be necessary. With these modifications, perhaps existing carrier relays can be depended upon to obtain fast and reliable tripping; or perhaps a high-speed back-up scheme should be included also. Assuming carrier, it is estimated that omission of the first zone impedance element may add from 0.5 to 1.0 cycle to the clearing time of the circuit breaker nearest the fault because of the longer operating time required by the carrier relays. Such an increase in relay time would be disadvantageous. It might be overcome by substituting high-speed current-balance relays for the discarded first zone impedance relay. The current-balance relays would provide high-speed sequential tripping and would back up the carrier relaying and vice-versa. Or after possible improvements in the relatively new phase-comparison carrier relaying scheme have been made, this scheme might prove satisfactory. For the present, the phase-comparison scheme is relatively slow and cannot be applied where the minimum 3-phase fault currents are only slightly greater than maximum load current, and therefore this scheme is not considered applicable.

With the foregoing considerations in mind, it was decided that a reasonably safe assumption of clearing time for use in the stability studies would be 3.5 cycles for the relay and circuit breaker nearest the fault and 7.0 cycles for the equipment at the far end. Such sequential tripping was used in the study, although with carrier relays, simultaneous tripping of both ends might result in clearing in about 4.5 cycles, which would produce slightly less shock to the system than the sequential tripping assumed.

No accurate data were available for time required to reinsert the capacitors in the circuit after clearance of a fault. Therefore, two values for reinsertion were assumed, either 7.0 or 10.5 cycles after the occurrence of the fault, and the amount of compensation required to obtain transient stability for each assumed time was determined. Thus a comparison is obtained showing the increase in compensation required for the same power transmitted when the reinsertion is delayed from 7.0 to 10.5 cycles.

Results and Discussion

Table II lists some of the results of a number of the studies made. For the first six cases listed, a constant percentage of compensation was used both before and after the faulted line section is removed, which assumes more compensation added

Table II. Summary of Stability Studies

Case Number		1	2	3	4	5	6	7	8	9
1. Constant per cent compensation. See item 5. (Before and after fault)	Initial per cent	60	63	63	65	66	75			
	Final per cent	60	63	63	65	66	75			
2. Constant ohmic value of compensation. (Before and after fault. No additional compensation added on removal of fault)	Initial per cent							100	100	0
	Final per cent							75	75	0
3. Compensation removed during fault		No	No	Yes	No	Yes	Yes	Yes	Yes	
4. Compensation reinserted after start of fault (cycles)				7.0		7.0	10.5	7.0	10.5	
5. Additional compensation added after start of fault (cycles). (Equally at both capacitor stations)		7.0	7.0	7.0	10.5	7.0	10.5			
6. Initial loading at receiver end (kw/1,000)		311	324	324	325	314	315	324	324	150
7. Initial operating angle between E_{is} and E_{ir} (degrees)		32.5	34	34	34.5	29	24.5	12.5	12.5	28.5
8. Maximum angle of swing (degrees)		102	91	Unstable	Unstable	91	98	102	Unstable	91.5
9. Shunt reactor (kilovars/1,000)*						109	113	114	114	

* Installed at point 1′ (see Figure 1) to reduce maximum voltage at point 1 and to neutralize a portion of line charging current and decrease current flowing in section S-1 below thermal limit.

Fault type, location, and clearing time: Two phase to ground at point 1 (see Figure 1). Near end cleared in 3.5 cycles, far end in 7.0 cycles.

Table III. Net Acceleration Ratios Obtained. Fault at Point 1

Case No.	Initial Per Cent Compensation	Generator Rating, Kva/1,000		Initial Power Output, Kw/1,000		*Accelerating Power, Kw/1,000		*Acceleration, Degrees per Second^2			*Acceleration Ratio Case 9 = 1.00
		S	R	S	R	S	R	S	R	Net	
1†	60	375	640	359	634	229	172	1,645	581	1,064	0.97
2†	63	375	640	376	624	241	172	1,735	581	1,154	1.05
3	63	375	640	375	618	241	−28	2,020	−95	2,115	1.93
4†	65	375	640	375	618	240	166	1,730	561	1,169	1.07
5	66	375	640	373	604	273	−32	1,965	−108	2,073	1.89
6	75	375	640	367	600	274	−24	1,970	−81	2,051	1.85
7 and 8	100	375	640	377	590	302	−20	2,175	−68	2,243	2.05
9	0	185	370	164	390	113	95	1,650	554	1,096	1.00
10‡	0	185	370	164	390	152	147	2,215	858	1,357	1.24‡

* At instant fault is applied.

† Capacitors not short-circuited during fault; for other cases, capacitors short-circuited during fault.

‡ 3-phase fault applied; for all other cases 2-phase-to-ground fault applied.

after the faulted line section is cleared. This assumption was made to determine if there were any appreciable increase in power limit if a portion of the compensation is added after the fault is removed. Such addition of compensation at all capacitor stations after the fault is cleared may introduce another problem because of the distance between stations. A transfer tripping scheme using additional carrier channels probably would be needed to obtain the high speed of operation that would be necessary.

In some of the cases listed the compensation is removed completely during the fault; in others it is left in the circuit. In cases 7 and 8 a constant ohmic value of compensation is used before and after the faulted section is cleared. With constant ohmic value, the per cent compensation is decreased when the faulted line section is removed from service. For comparison purposes, case 9 shows results for the system of Figure 1 but without any compensation, so that the power delivered has to be reduced materially (to approximately one-half) to maintain stability as compared to the compensated systems. Proportionately reduced size of terminal equipment corresponding to the reduced power transmitted is used in case 9. Table II also lists the power delivered for each case and the initial operating angle before the fault is applied. In some cases a shunt reactor is included at point $1'$ in Figure 1, which was placed there to keep the line voltage at point 1 from being excessive and to reduce the line-charging current flowing through the sending-end section, thereby increasing the delivered power without exceeding the thermal current-carrying capacity of the conductor when one line section is out of service. Incidentally, the reactor also increases the stability of the system.

A comparison of cases 1 and 5 of Table II indicates that approximately 10-per-cent increase in capacitor impedance, voltage, and kilovolt-ampere capacity is needed, or an increase of 6 per cent of line compensation is needed when the capacitors are removed from service during the fault period of 7.0 cycles. A comparison of cases 5 and 6 shows that the extra 3.5 cycles of delay in getting the capacitors back into service after the fault is cleared require another 13.5-per-cent increase in capacitor kilovolt-amperes or 9-per-cent increase in line compensation. Both of these effects require from 20- to 25-per-cent increase in capacitor kilovolt-amperes, or about 15-per-cent increase in line compensation over that needed if the capacitors were not taken out of the circuit for the 10.5 cycles.

For 325,000 kw delivered and on the basis of constant capacitor ohms (case 7), 75-per-cent final line compensation or 100-per-cent initial compensation (when only two capacitor stations are employed) is required when the capacitors are removed from service during the fault and are reinserted at the time the fault is cleared. It will be observed that the initial operating angle between the internal voltages of sending and receiver machines decreases with increasing compensation. Such a decrease in initial operating angle apparently does not change materially the amount of power which can be transmitted with stability, provided the final percentage compensation is identical for two otherwise similar cases. Hence, there is no real advantage from a stability standpoint of introducing the complication of switching in a portion of the compensation at each capacitor station after the fault is removed.

Figure 2A is a load-flow diagram of the conditions of initial loading for the system of Figure 1 using 75-per-cent initial compensation (case 6). The figure is included to show the various bus voltages and reactive power flow through the various lines and capacitor stations. Figure 2B is a vector diagram of the system under these loading conditions. It is of special interest to observe the relative phase angles obtained in machines, lines, and series capacitors. Figure 3 is a vector diagram of the system after the faulted line section between points S and 1 has been removed, the capacitors reinserted, and the system has swung to the maximum phase angle. Note the large phase angle, 117 degrees, between E_{ts} and point 1, and the corresponding low values for E_s, 66 per cent of normal, and for E_2', 47 per cent of normal. The current through the remaining line section has increased from its initial value of 450 amperes to 1,444 amperes or 3.26 times the initial value. It is also of interest to observe that the voltage across the capacitors at station 1 is 155 per cent of rated value.

It is desirable to compare the relative net accelerations between machine rotors at the instant the fault occurs for the various cases studied. Table III lists the net accelerations assuming double-line-to-ground faults for compensated lines with and without compensation removed. For further comparison, acceleration for completely uncompensated lines (case 9) and for a 3-phase fault on uncompensated lines (case 10) is included. Using the net acceleration for the uncompensated line and a double-line-to-ground fault (case 9) as a standard of comparison, it will be observed that for cases 1, 2, and 4 in which the compensation is not removed, the acceleration ratio varies from 0.97 to 1.07. In other words, the shock to the system is almost directly comparable with case 9. However, when the compensation is short-circuited during the fault, the acceleration ratio varies from 1.85 to 2.05, or the shock to the system is approximately double that for an uncompensated line as for a system in which the capacitors are left in during the fault. For the uncompensated system and a 3-phase fault (case 10) the shock to the system is only 24 per cent greater than the double-line-to-ground fault in case 9. These relative accelerations are an indication why additional compensation is needed in order to maintain stability in cases where the compensation is removed during the fault.

As further comparison, sets of net acceleration-angle curves are shown in Figures 4A and 4B. Figure 4A shows acceleration-angle areas for case 2 in which the compensation was not removed during the fault. With respect to these curves, it can be shown that the positive

Figure 48. Net acceleration between machine rotors for 230-kv 400-mile compensated system (case 6)

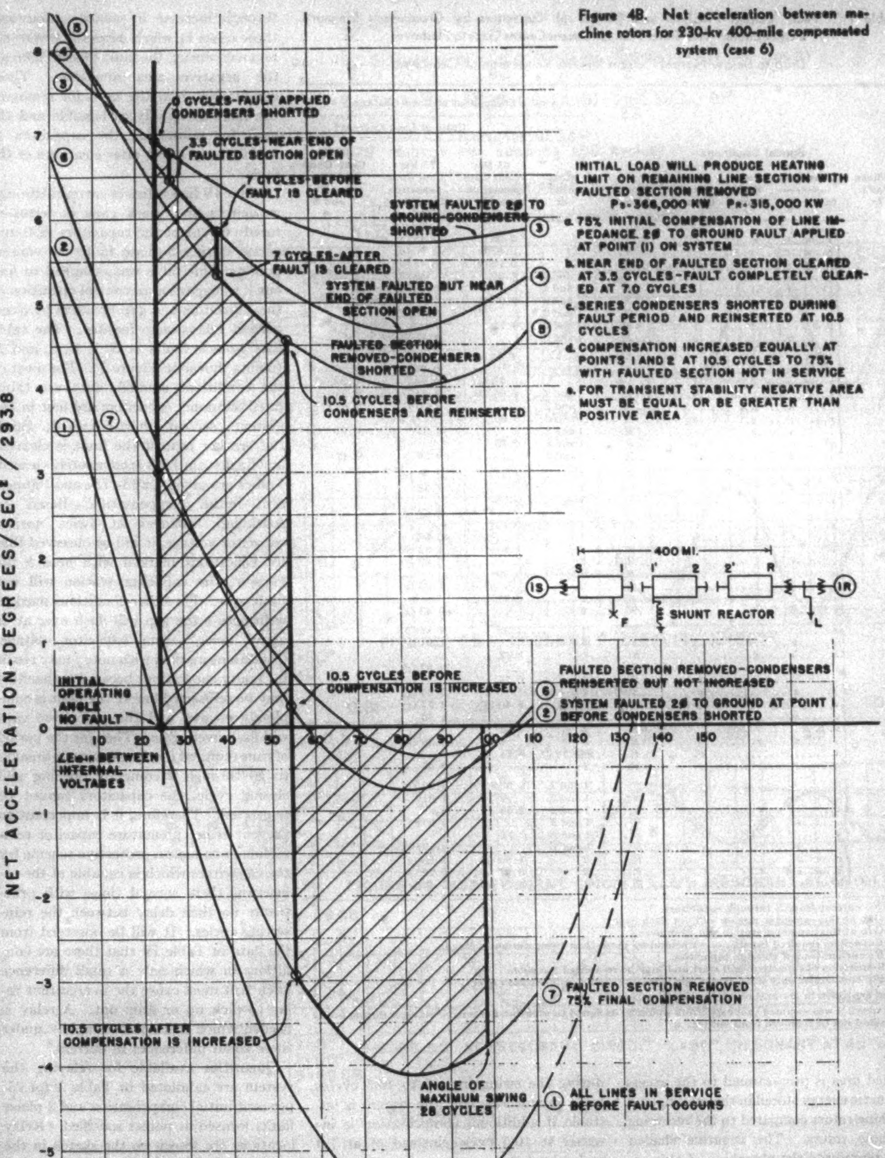

Table IV. Fault Current Analysis and Control of Capacitors by Overcurrent Elements. Assumptions: Two Times Normal Current Causes Gaps to Flashover

Drop to Below Normal Current Initiates Reinsertion of Capacitors

	Special Conditions						Component of Load Current		
							Units of Load Current*		
3-Phase Fault At	Capacitors In	Capacitors Out	Near End Open	Fault Cleared	Supplied From	Measured at	63 Per Cent Compensation Case No. 2†	75 Per Cent Compensation Case No. 6	100 Per Cent Compensation Case No. 7 and 8
S	x or	x			S	S	2.26 (a)	2.45 (a)	
	x or	x	x		S	S	0.61 (a)	0.56 (a)	
	x or	x		x	S	S	0.76 (a)	0.45 (a)	
	x				R	1 and 2	1.20	1.38 (b)	
		x			R	1 and 2		0.63 (g)	
	x		x		R	1 and 2	0.91		
		x	x		R	1 and 2		0.48 (g)	
	x			x	R	1 and 2	0.72		
		x		x	R	1 and 2		0.47 (g)	
1	x or	x			S	S	1.24 (a)	1.34 (a)	1.26 (a)
	x or	x	x		S	S	1.01 (a)	0.99 (a)	0.87 (a)
	x or	x		x	S	S	0.76 (a)	0.54 (a)	0.37 (a)
	x				R	1 and 2	2.05 (d)	3.14 (c)	3.99 (c)
		x			R	1 and 2		0.90 (f)	0.85 (f)
	x		x		R	1 and 2	0.66		
		x	x		R	1 and 2		0.37	0.30
	x			x	R	1 and 2	0.73		
		x		x	R	1 and 2		0.56	0.47
1′	x				S	1	1.97	2.37 (c)	
		x			S	1		1.38	
	x		x		S	1	0.64		
		x	x		S	1		0.60 (f)	
	x			x	S	1	0.72		
		x		x	S	1		0.53	
	x				R	2	1.24	1.32 (b)	
		x			R	2		0.87 (g)	
	x		x		R	2	0.99		
		x	x		R	2		0.63 (g)	
	x			x	R	2	0.71		
		x		x	R	2		0.43 (g)	
2	x				S	1	1.22	1.46 (b)	
		x			S	1		1.01 (g)	
	x		x		S	1	1.02		
		x	x		S	1		0.82 (g)	
	x			x	S	1	0.72		
		x		x	S	1		0.53 (g)	
	x				R	2	2.40 (d)	2.77 (c)	
		x			R	2		1.33	
	x		x		R	2	0.62		
		x	x		R	2		0.39 (f)	
	x			x	R	2	0.71		
		x		x	R	2		0.43	
2′	x				S	1 and 2	1.80	2.37 (c)	
		x			S	1 and 2		1.01	
	x		x		S	1 and 2	0.63		
		x	x		S	1 and 2		0.50 (f)	
	x			x	S	1 and 2	0.71		
		x		x	S	1 and 2		0.50	
	x or	x			R	R	1.30 (e)	1.28 (e)	
	x or	x	x		R	R	0.97 (e)	0.79 (e)	
	x or	x		x	R	R	0.74 (e)	0.45 (e)	

(a)—No current from S through capacitors.
(b)—Gap at this capacitor station will not flash over.
(c)—Gap at this capacitor station will flash over.
(d)—Capacitors required capable of withstanding more than twice normal voltage.
(e)—No current from R through capacitors.
(f)—Reinsertion of capacitors will start and may be reinserted too soon.
(g)—For comparison only since gap at this capacitor station will not flash over.

* Load expressed in per unit of full load (900 amperes).

† For case 2 it was assumed that capacitors would be designed to withstand excessive voltages during faults and would not be removed from service.

shaded area is proportional to the excess of kinetic energy stored in the sending-end machine rotors compared to the receiving machine rotors. The negative shaded area represents the absorbing of this excess kinetic energy. The system is stable provided the negative area is greater than the positive area. A comparison of positive area produced during the swing from 0 to 7 cycles (case 2) to that produced during the swing from 0 to 10.5 cycles (case 4) indicates why the system is unstable if additional compensation is inserted at 10.5 cycles instead of at 7.0 cycles.

Figure 4B shows acceleration curves for case 6 in which the compensation is removed during the fault. A much larger positive area is produced by removal of the compensation. It must be offset through increase in compensation over those cases in which capacitors were not removed during the fault so as to increase the negative area available. These curves also show the need for removing the fault as rapidly as possible and the need for reinserting the capacitors as rapidly as possible after clearance of the fault.

Table IV lists ratios of currents through capacitors and spark gaps in terms of rated current of the capacitors (900 amperes) during 3-phase faults at various points. This table was compiled to find out if a suggestion to control operation of the capacitors and gap breakers by overcurrent relays were feasible. The table lists 3-phase faults at S, 1, 1′, 2, and 2′ (points shown on Figure 1). For most of the conditions studied, data are tabulated assuming capacitors are first in the system; second, short-circuited; third, the breaker nearest the fault is cleared; and finally, the fault is completely cleared. Values are given for 63-, 75-, and 100-percent initial compensation. Based on spark-gap flashover at twice normal capacitor voltage, it will be observed that for conditions marked with note *b* the gaps at the capacitor station will not flash over. For other conditions marked with note *c* the gap will flash over at or above twice normal capacitor voltage. Conditions marked with note *f* may result in faulty operation because capacitors may be reinserted before the fault is completely cleared and the protective gaps will flash over again. Then for the period of time required for the gap circuit breaker to go through a complete opening and closing cycle, the capacitors cannot be reinserted. Therefore, it is important to prevent either premature capacitor reinsertion or to have a protective scheme for the capacitors which is capable of the reinserting them several times with practically no time delay between the reinserting cycles. It will be observed from the data in Table IV that there are conditions in which only a small difference in current must cause the overcurrent relay to pick up or drop out. A relay is needed which will operate rapidly under these small differences in current.

Quantities available for relaying the system are tabulated in Table V for 75-per-cent initial compensation and 3-phase faults located at points specified. Relay locations are shown on the sketch in the table. Voltage, current, and impedance, as seen by the relay, are tabulated. One relaying problem which may result in faulty operation of directional relays at points K and M occurs for a fault near point 1. Both relay elements should

Figure 5. Starting transient in an *L-R-C* series circuit

$$i = \frac{E_m}{Z}\,\mathrm{Sin}(\omega t + \lambda - \theta) + I_t\,\epsilon^{-\alpha t}\,\mathrm{Sin}(\beta t - \sigma)$$

$$\text{or } i = \frac{E_m}{Z}\,\mathrm{Sin}(\omega t + \lambda - \theta) + \frac{E_d}{\beta L}\,\epsilon^{-\alpha t}\,\mathrm{Sin}\,\beta t - \frac{E_m}{Z}\,\mathrm{Sin}(\lambda - \theta)\,\epsilon^{-\alpha t}\,\mathrm{Cos}\,\beta t$$

$$\text{Where } E_d = E_m\,\mathrm{Sin}\,\lambda - \frac{Q_0}{C} - \frac{E_m \omega L}{Z}\,\mathrm{Cos}(\lambda - \theta) - \frac{E_m R}{2Z}\,\mathrm{Sin}(\lambda - \theta)$$

$Q_0 = 0$ charge on condenser at $t = 0$

$\omega = 377$ radians per seconds (60 cycles)

$\alpha = \frac{R}{2L}$ λ = angle of E at $t = 0$

$\beta = \sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}}$ θ = impedance angle

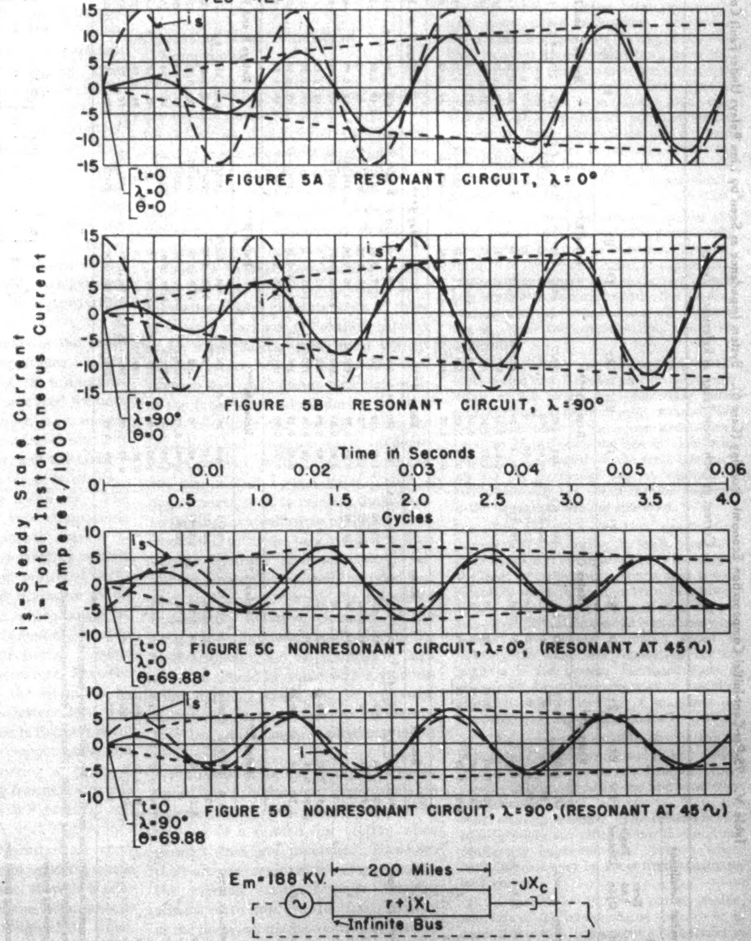

Fig. No.	r	jX$_L$*	-JX$_C$*	% Neutralized
5a & 5b	12.9 Ω	82 Ω	82 Ω	100%
5c & 5d	12.9 Ω	82 Ω	46.5 Ω	57%

* At 60 CYCLES

Table V. 75-Per-Cent-Initial-Compensation Economic Loading (Case 6). System Impedance as Seen by Line Relays Under Fault Conditions

Current at Each Point Taken as Reference Vector

3-Phase Fault At	Special Conditions				I*	E‡	Z†	I*	E‡	Z†	I*	E‡	Z†	I*	E‡	Z†	I*	E‡	Z†	I*	E‡	Z†
	Capacitors In	Capacitors Out	Near End Open	Fault Cleared																		
					Relay B**			Relay C**			Relay D***			Relay E***			Relays F and H			Relays K and M		
Initial loading, amperes, kv, ohms					442	240/0	542/0	−433	236/−8.5	−544/−8.5	442	240/0	542/0	−433	236/−8.5	−544/−8.5	416	240/−4.5	579/−4.5	406	240/9.5	592/9.5
(a)S	x				6.24	0	0	1.44	0.53	202/80	−1.41	0	0	1.44	0.53	202/80	−1.49	0.06	−29/−13	−1.48	0.15	−60/−17.5
		x			5.46	0	0	0.66	0.25	203/81	−0.65	0	0	0.66	0.25	203/81	−0.77	0.24	−183/81.5	−0.69	0.49	−416/81
		x	x		0			1.53	0.54	193/83	1.14	0.88	420/55	−1.17	0.54	−253/34	0.63	0.53	488/−50.5	0.53	0.88	746/−55
		x		x	0			0			0.91	1.02	607/18	−0.99	0.91	−502/12	0.52	0.89	996/−20	0.48	0.92	1,133/−7
(b)1	x				1.37	0.50	197/81	7.89	0	0	1.37	0.50	197/81	−1.40	0	0	−2.99	1.22	−234/−87	−3.24	1.32	−240/−81.5
		x			1.37	0.50	199/81.5	3.00	0	0	1.37	0.50	199/81.5	−1.40	0	0	−0.96	0	0	−0.96	0.33	−203/81.5
		x	x		1.96	0.69	192/82.5	0			0.74	0.69	504/5	−0.76	0.73	−523/−17.5	0.50	0.72	821/−44	0.40	0.80	1,196/−36
		x		x	0			0			0.91	1.02	607/18	−0.99	0.91	−502/12	0.52	0.89	996/−20	0.48	0.92	1,133/−7
					Relay F**			Relay G**			Relay H***			Relay J***			Relays C and E			Relays K and M		
Initial loading, amperes, kv, ohms					416	240/−4.5	579/−4.5	−406	241/−10.5	−594/−10.5	416	240/−4.5	579/−4.5	−406	241/−10.5	−594/−10.5	−433	236/−8.5	−544/−8.5	406	240/9.5	592/9.5
(c)1'	x				6.25	0	0	1.47	0.51	205/82	−1.43	0	0	1.47	0.51	205/82	−2.46	0.96	−212/−88	−1.47	0.07	−29/−16
		x			3.71	0	0	0.96	0.32	200/82.5	−0.94	0	0	0.96	0.32	200/82.5	−1.44	0	0	−0.96	0.32	−200/82.5
		x	x		0			1.74	0.56	193/82.5	0.93	0.77	476/38	−0.95	0.56	−352/15	−0.62	0.78	−682/56	0.70	0.57	477/−59
		x		x	0			0			0.95	0.93	563/1	−0.96	0.91	−563/−7	−0.55	0.94	−936/32	0.48	0.91	1,130/−7
(d)2	x				1.43	0.50	202/81	7.50	0	0	1.43	0.50	202/81	−1.46	0	0	−1.52	0.11	−38/−50	−3.07	1.09	−210/90
		x			0.99	0.34	200/81.5	3.73	0	0	0.99	0.34	200/81.5	−1.01	0	0	−1.05	0.35	−180/82	−1.48	0	0
		x	x		1.70	0.57	193/84	0			0.85	0.57	385/−23	−0.87	0.74	−504/−47	−0.85	0.58	−370/53	0.43	0.74	1,011/−47
		x		x	0			0			0.95	0.92	561/0	−0.96	0.91	−565/−7	−0.55	0.94	−932/31	0.48	0.92	1,134/−7
					Relay K**			Relay L**			Relay M***			Relay N***			Relays C and E			Relays G and J		
Initial loading, amperes, kv, ohms					406	240/9.5	592/9.5	−414	221/3	−535/3	406	240/9.5	592/9.5	−414	221/3	−535/3	−433	236/−8.5	−544/−8.5	−406	241/10.5	594/−10.5
(b)2'	x				6.07	0	0	1.39	0.53	204/81	−1.42	0	0	1.39	0.53	204/81	−2.46	1.00	−222/−83.5	−2.54	0.91	−213/−89.5
		x			3.28	0	0	1.39	0.52	202/82	−1.42	0	0	1.39	0.52	202/82	−1.05	0.35	−180/81.5	−1.05	0	0
		x	x		0			2.01	0.73	193/81.5	0.94	0.83	526/27.5	−0.92	0.73	−423/6	−0.51	0.92	−969/42	−0.47	0.83	−1,052/27.5
		x		x	0			0			0.96	0.95	575/9	−0.98	0.96	−525/2.5	−0.62	0.99	−1,035/27	−0.49	0.95	−1,151/9
(a)R	x				1.53	0.52	199/81.5	7.67	0	0	1.53	0.52	199/81.5	−1.50	0	0	−1.53	0.17	−80/−33	−1.53	0.07	−29/−22.5
		x			0.82	0.28	199/81	7.15	0	0	0.82	0.28	199/81	−0.80	0	0	−0.84	0.54	−352/82.5	−0.82	0.27	−199/81
		x	x		1.67	0.55	193/82	0			0.90	0.55	358/−35.5	−0.88	0.84	−507/−58	−0.70	0.71	−551/55	−0.66	0.54	−488/40
		x		x	0			0			0.98	0.95	574/10	−0.98	0.96	−524/2	−0.62	0.99	−1,035/27	−0.49	0.95	−1,150/10

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* Negative denotes current in nontripping direction. Per unit of initial loading current.

† Negative Z denotes current in nontripping direction. Negative θ denotes capacitive impedance.

‡ Per unit of initial operating voltage.

** Relays on faulted line.

*** Relays on line paralleling faulted line.

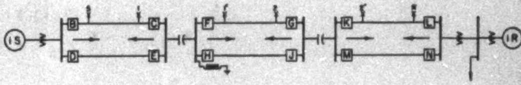
(a)—Neither gap will flash over.

(b)—Both gaps will flash over.

(c)—Gap at capacitor station 1 will flash over. Gap at capacitor station 2 will not flash over.

(d)—Gap at capacitor station 1 will not flash over. Gap at capacitor station 2 will flash over.

Note: Table gives conditions existing on the system assuming both capacitor gaps flash over before the first circuit breaker opens. Actually such will not always be the case. See footnotes (a)–(d).



block tripping for this fault. Before the gaps flash over, relays *K* and *M* see a high capacity reactance with voltage lagging the current by over 80 degrees in the blocking direction. After the gaps flash, the voltage leads the current by over 80 degrees (in blocking direction). However, these relays should trip for a fault near the receiver bus in which case the voltage is leading the current by over 80 degrees (tripping direction). Hence, this relay should have a torque–phase angle characteristic which will cause it to trip for the voltage leading the current by 80 degrees, but it must block for the voltage leading the current by 100 degrees. It is believed that existing standard directional elements will need to be improved to obtain both high-speed and positive contact closing when operating so near the zero torque characteristic curve. Similar requirements must be met by relays *F* and *H* for a fault near point 1. Additional studies will be needed to determine the effects of unbalanced faults on this phase shift. Also the effects of unbalanced impedances in case individual spark gaps are used for each phase of the compensating capacitor must be studied.

The effects of transients in the series *L-R-C* circuit are of importance. When such a circuit is energized, the instantaneous current i which is produced includes two components, a damped a-c transient which oscillates at the natural or resonant frequency of the circuit, and the steady-state a-c wave which is forced to oscillate at the 60-cycle supply frequency. The summation of these two components produces the instantaneous current i. Figures 5A and 5B show starting current and steady-state current projected backward to zero time for resonance at 60 cycles and for λ, the phase angle at which the voltage first is applied, equal to 0 and 90 degrees, respectively. Figures 5C and 5D show similar currents when the natural frequency of the circuit is 45 cycles. Assuming no system load initially, the system shown in Figure 1 when using 75-per-cent line compensation and assuming a fault occurring at point 1′ would produce starting transient currents similar to those shown in Figures 5C and 5D.

The transient and steady-state components of current are in opposing phase at the beginning; hence, there is a relatively slow build up of the instantaneous current to the final steady-state value. Another way of stating this phenomena is that it takes time to store energy in the dielectric and magnetic fields through the resistance of the circuit; hence, the current is larger the second half cycle than it is the first half cycle. The time constant of this build up is from 2.5 to 3.5 cycles, which is of course appreciable in terms of gap flashover time and of operating time of high-speed relays. A factor which may be important from a directional relaying standpoint (Figures 5C and 5D), is the shift in phase angle during the first two cycles of the total current wave i as compared to the steady-state wave. These factors must be investigated further with respect to their effects on relays.

Operating Requirements of Equipment

With respect to the operating requirements and design of control equipment for capacitors, it is believed that highest speed of operation cannot be obtained by use of three pieces of conventional equipment, namely, relays, a circuit breaker or contactor, and a spark gap. Instead, the functions of this equipment should be combined in fewer pieces of apparatus.

From a system performance standpoint the ideal arrangement might have somewhat similar characteristics to those of a rotary spark gap or lightning arrester which would flash over when the voltage became greater than twice normal, continue to flash over during each half cycle of the fault, and seal off during the next half cycle after the voltage across the capacitor had dropped to a predetermined amount, perhaps normal value or 150 per cent of normal value. Such a device might tend to reinsert the capacitor during each half cycle of the fault but would fail to do so because of the high voltage obtained. However, it would fulfill the important operating condition that the capacitor would be reinserted immediately following clearance of the fault. Investigation of oscillations in the system which might be produced by such a device should be made with a view toward damping them out, or otherwise minimizing any system or apparatus troubles which such oscillations might produce.

Protection of the line sections can be obtained by a high-speed current-balance scheme which will provide sequential tripping of a faulted line section when parallel lines are involved. However, there are good reasons for trying to modify the existing standard-carrier relaying scheme or to improve the characteristics of the phase-comparison carrier scheme so that one or the other can be used with series compensation. With respect to developing a directional relay characteristic to be used with the standard carrier scheme, it may be possible to combine the capacitor voltage with the regular line voltage and current to obtain proper phase-angle characteristics. This seems to be promising since it is the capacitor voltage which causes the trouble in the first place.

Conclusions

The results of this study indicate

(*a*). That compensation is very effective in increasing the allowable line loading.

(*b*). Building the capacitors to withstand the high voltages during short circuits in most cases cannot be justified. However, for system arrangements having relatively high terminal impedance at both ends and a relatively low percentage of line compensation (produces a relatively low ratio of maximum fault current to maximum load current), design *A* which does not remove the capacitors from service during a fault may have distinct advantages over design *B* and, therefore, although a more expensive installation, may be economically feasible where firm power is a consideration.

(*c*). If capacitors are removed during the fault period of seven cycles, approximately ten per cent increase in capacitor kilovolt-amperes or six per cent increase in line compensation is required for the particular system studied to obtain the same line loading with stability, as compared to system quantities in which the compensation is not removed.

(*d*). If capacitors are removed during the fault period of 7.0 cycles plus an additional delay of 3.5 cycles for reinserting them after the fault is cleared, approximately 25-per-cent increase in capacitor kilovolt-amperes or 15-per-cent increase in line compensation is required for the same line loading on the particular system studied, as compared with the compensation required if the capacitors are not removed.

(*e*). Due to high machine accelerating powers produced during the fault, especially if the capacitors are removed, fast clearing of the fault when compensation is used is considerably more important than in conventional systems. Fast reinsertion of capacitors after the fault is cleared is extremely important. To obtain increased speed in clearing the fault and reinserting the capacitors in the system, increased costs of relays and circuit breakers may be justified. Similarly, higher inertias of machines are desirable and may be economically justified.

(*f*). Assuming that modification of the existing standard relay schemes is not practical, a highspeed current-balance relay scheme appears to fulfill the line protection requirements more completely than any other existing scheme. The relatively new phase-comparison schemes are slow and not completely applicable on systems where fault currents may be lower than maximum load current.

(*g*). Considerable further study, design, and test of special apparatus seem to be in order to obtain equipment characteristics which will meet best the new problems encountered when series compensation is used

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Maintenance of Rectifiers on Electrochemical Installations

Discussion of paper 46-96 by J. E. Housley and G. N. Hughes, presented at the AIEE Southern District meeting, Asheville, N. C., May 14–16, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 436–41.

H. G. Harvey and **T. J. Woth** (Westinghouse Electric Company, Atlanta, Ga.): Industry often does not have the opportunity to benefit from such a wealth of experience that has been obtained during the past few years at the Aluminum Company of America plant and as described by J. E. Housley and G. N. Hughes. The experience gained by the operation of such a large number of units varying in type, construction, and circuit design, most certainly should serve as a valuable guide in future installations.

The designer, to a large extent, depends upon operating problems to guide him in his development and research activities. If we can look forward to receiving such a valuable operating report periodically, and I believe we can, we should feel confident that the rapid progress that has been made during the past years in electronic conversion equipment will continue.

During the war, the installation of rectifiers other than electrochemical declined somewhat. However present indications are that rapid expansion will be made in the use of these devices, especially to supply, street railway, trolley buses, and mining loads. Although the load factor involved in these applications probably will average as low as 25 per cent as compared to more than 90 per cent for the electrochemical applications, the majority of these units will be of single-anode, multiple-tank continuously-pumped type. Operating experience with a limited number of ignitron rectifiers installed before the war, operating under the rapidly varying railway load conditions, indicates that highly successful operations can be expected.

Recent experimental work in connection with artificial arc-backs has cast some new light on the arc-back problem and should be helpful in still further reducing this trouble in the future.

The main cause of arc-backs, as described by the authors, is assumed to be the formation of a cathode spot on the anode. However, experimental work during the development of the ignitron principle indicated that if the source of ionized vapor could be eliminated entirely during the negative half cycle, it should reduce the number of arc-backs greatly. This was accomplished by the use of the ignitor circuit which made it possible to eliminate entirely the separate d-c-excited keep-alive anodes. The absence of ionized vapor in the ignitron unit, except during the conduction period, undoubtedly accounts for the much lower arc-back rate for this type of unit.

All standard rectifiers are designed so that a voltage reduction of approximately 15 per cent can be obtained by delayed firing.

There are however a number of disadvantages which, as the authors have pointed out, make it desirable to operate normally with zero firing delay or as near to this point as possible. In addition to the increased tendency to backfire, and lower power factor, it should be pointed out that the magnitude of harmonic currents increases rapidly with the number of degrees of firing delay. Although this item may be of minor importance in the large multiphase installations, it is generally an important consideration in the smaller individual applications. In most of these cases, the high-voltage transformer taps can be arranged so that normal operation with zero firing delay will be satisfactory and the voltage control feature used only in emergencies or for the degassing operation.

The many helpful suggestions made by the authors in regard to special pressure tests for finding gas leaks, caution that should be used in making repairs, and critical performance during the degassing operation should be extremely helpful to the operators who now are planning initial installations of rectifier equipment.

Inductive Co-ordination Aspects of Rectifier Installations

Discussion of paper 46-105, an AIEE committee report, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 417–36.

R. D. Evans (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The contents of this paper should prove to be of great usefulness to the industry. In the preparation of this report special efforts have been made to provide a discussion of a method which could be used generally by application engineers of industrial organizations and power companies as well as by engineers of the electrical manufacturing and telephone companies. It is believed that this report is sufficiently self-contained and explicit to permit its use by a much wider group of engineers than was possible by previous technical reports on this subject.

The body of the report is basically simple in character, and this fact should not be disregarded by the presence of the appendixes which contain supplementary material of considerable interest and usefulness in special cases.

The report contains the term "commutating reactance," and the proposed standard for pool cathode mercury-arc power converters (AIEE Standard 6, June 1946) includes a definition of commutating reactance. A discussion of the meaning and significance of commutating reactance may be of interest.

For many years specialists in the application of rectifiers have used the term commutating reactance in all theoretical methods for estimating the performance of rectifier circuits from the standpoint of regulation and harmonics. Nevertheless, rectifier transformer nameplates in the past have not included commutating reactance, but instead have given another reactance which frequently has been referred to as nameplate reactance. The use of these two reactances of a rectifier transformer has been a source of confusion and to a considerable extent has limited the application of analytical methods for estimating harmonics in the circuits of rectifiers. This particular difficulty would be avoided in the future by the methods presented in this paper and in the treatment included in the proposed rectifier standards.

Nameplate reactance of a rectifier transformer may be obtained by applying polyphase voltage to the primary of a rectifier transformer with all secondary terminals short-circuited. Commutating reactance of a rectifier transformer is half of the reactance in the circuit involving the anodes between which commutation takes place. Commutating reactance may be measured by applying single phase voltage to the rectifier transformer terminals between which commutation takes place with the a-c line terminals of the rectifier transformer short-circuited. The two reactances may be converted from primary to secondary voltage bases, or vice versa, by using the well-known conversion methods. It is important, however, to recognize that these two reactances for a particular rectifier transformer will have, in general, different numerical values. These differences arise from the circumstances which cause the mutual effects of the windings to be different for the two circuit conditions. Thus the difference between nameplate reactance and commutating reactance depends in part upon the circuit connections and in part upon winding construction which affects the mutual inductance between the windings. Because of these circumstances it is not feasible to provide a simple equation relating the numerical values of these reactances. In the future the commutating reactance will be given on the nameplates of rectifier transformers. For existing transformers this information can be obtained by tests in the field or may be supplied by the manufacturer of the transformer who usually makes such a test in the factory or can estimate the reactance from design proportions. Nameplate reactance is of use principally in connection with the calculation of maximum short-circuit current in the rectifier transformer.

Protection of Powerhouse Auxiliaries

Discussion and closure of paper 46-169, an AIEE committee report, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 746–51.

W. R. Brownlee (The Commonwealth and Southern Corporation, Jackson, Mich.): While this report is based on a relatively small number of replies, it is believed that a representative cross section of experience has been secured. On many aspects of the problem a majority concurrence is indicated, but there are a number of cases of sharp divergence in practice. It is felt that this also holds true when a large number of cases are considered. For example, most users omit undervoltage protection, although some have retained it. Many connect motor thermal overload elements to operate an alarm for essential motors, but several connect the thermal devices to trip the motors. A large proportion of users rely on transformers for station service supply, but a few utilize shaft generators or even house turbines.

In order to provide the reliability which it generally is agreed is so necessary in station auxiliary performance, it is believed that close attention is required to these points

1. Eliminate all (or almost all) undervoltage protection even of the time delay type on all essential motors.

2. Connect thermal relay elements to provide alarm or indication for essential motor circuits, using elements with generous settings to trip nonessential motors only.

3. Provide station auxiliary sources from transformers, preferably connected at generator voltage, with proper co-ordination of control and protection.

Several factors have contributed to the trend toward these principles in the treatment of station auxiliaries. Undervoltage protection was at one time required because of the use of compensator starting or slipring induction motors. With modern designs all essential motors may be of the squirrel-cage line-starting type so that undervoltage protection is not required for this purpose. The desirability of safety interlocking was responsible for the use of undervoltage protection in the past, but recent plants make use of safety devices responding to the driven quantity, such as differential air pressure relays in the exhauster pipes, and time lagged pressure relays in the induced draft fan systems. Similarly the present day practices in central stations differ from the typical industrial plant in that special attention is given to the co-ordination of motors with the drive characteristics, maintenance is performed on a program basis, and close surveillance by station operators eliminates many causes of motor overloading.

With a co-ordinated system of station auxiliaries as outlined in a foregoing paragraph, a highly reliable and efficient system may be provided. Tests on such a co-ordinated scheme have been described[1] showing that the system is immune to external short circuits since a total interruption of auxiliary power for as much as five seconds does not interfere with boiler and turbine performance. Test also have shown that in severe emergencies voltage may drop as low as 60 per cent of normal for a sustained interval without loss of flame or reduction in capacity. With such a system of electrical auxiliaries, the margin for emergencies in boiler and turbine operations is inherently greater than can be provided in the interconnected transmission system to withstand extreme voltage reductions, so that the over-all system emergency capability need not be limited because of station auxiliaries.

REFERENCE

1. ELECTRIC DRIVES FOR STEAM-ELECTRIC GENERATING-STATION AUXILIARIES. Wm. R. Brownlee, J. A. Elzi. AIEE TRANSACTIONS, volume 64, 1945, November section, pages 741–5.

E. L. Michelson (sponsor, relay subcommittee, AIEE committee on protective devices): W. R. Brownlee's discussion emphasizes the desirability of simple and reliable relaying methods for powerhouse auxiliaries. It is necessary for the auxiliary power system to be more reliable than any other component of the power system, because of the greater hazard which accompanies the loss of auxiliary power in the generating station.

The recent trend toward the use of motor driven exciters further increases the need of reliable system of auxiliary power. Since the loss of supply to the exciter set for even a brief interval may result in the loss of excitation to the main generating unit, continuity of service to the exciter set is even more important than the supply to other auxiliary services.

Brownlee mentions that the use of undervoltage relays largely has been eliminated for the protection of auxiliary power motors. It should be noted that these relays are being used in many installations for the purpose of restoring service to an auxiliary power bus which has lost its normal source of supply.

Mica Capacitors for Carrier Telephone Systems

Discussion of paper 46-151 by A. J. Christopher and J. A. Kater, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 670–4.

N. N. Babcock (Western Electric Company Kearny, N. J.): As mentioned in the paper by Christopher and Kater silvered mica capacitors have been made for many years, but only in the past decade with the development of new materials and methods have they proved desirable from either a design or a manufacturing standpoint. The first metalized mica capacitors were made basically by one of three processes.

1. Chemically depositing silver or other metal.
2. Cathode sputtering.
3. Evaporation of metals in a vacuum.

In general, such capacitors, in addition to being quite expensive to manufacture, tended to be unstable with variable voltage and temperature[1] and to possess undesirable frequency characteristics caused chiefly by the thinness and discontinuity of the coatings serving as the electrodes.

The application of metallic coatings, especially those of gold and silver, to ceramics by the method of firing on a glasslike matrix containing the finely divided metal or its salts is an ancient art. Leyden jars have been made in this manner for a long time. The extension of the method to mica capacitors seemed obvious, but the china decorators' pastes fired at temperatures too high to be tolerated either electrically or mechanically by the mica. Anticipating that many advantages might be gained, the Western Electric Company initiated a development about ten years ago which, in co-operation with commercial manufacturers of ceramic decorators' pastes, resulted in improved silver pastes firing at lower temperatures and possessing the desirable characteristics of good adherence to mica, ease of application, and controllable thickness of coating which now form the basis of the silvered mica capacitor industry.

The silver solutions now used by the Western Electric Company have been developed particularly for spray application. The silver is in finely divided metallic form with carefully controlled amounts of low melting point glass of special composition and an extender suspended in a suitable vehicle. Firing of the silver coating after spraying on the mica plates is accomplished in a well-ventilated electrically heated conveyor furnace. The electrical properties of the silvered mica plates or laminations have been found to be affected noticeably if the composition of the silver formulation is not controlled closely.

The assembly of the units, whether intended for use in the potted types or the molded phenolic encased capacitors, is essentially the same. In the case of the molded types the relatively small effect of pressure on the silvered mica unit allows a reliable prediction to be made of the capacitance rise that will take place during the molding operation. Consequently in commercial production capacitance tolerances of plus or minus the quantity (0.5 per cent plus one micromicrofarad), without resorting to selection methods, regularly are obtained. Three sizes of the molded type capacitor with wire leads are manufactured. The capacitance range covered by this group is two micromicrofarads to 0.02 microfarad. For the potted types, in which the silvered mica unit is placed in an extruded aluminum can filled with wax, there are three basic sizes of mica laminations used. From these basic unit sizes five different sizes of finished capacitors may be built up. The cans themselves may carry mounting lugs or not, depending upon the end application of the capacitor.

The testing equipment used in the manufacture and final inspection of these capacitors necessarily is designed and calibrated to a high order of accuracy. The capacitance and conductance of every capacitor is checked carefully in final inspection on precision bridges. The dielectric strength and insulation resistance requirements are checked on 100 per cent of the product. Temperature coefficient of capacity, drift, and quality of seal against the effects of humid atmospheres are checked on a sampling inspection basis.

It has been found advantageous at the Western Electric Company to carry out all assembly operations on silvered mica capacitors and the final testing as well, in an atmosphere of not over 45 per cent relative humidity and at a temperature held close to 75 degrees Fahrenheit. The adverse effects of adsorbed moisture on the mica laminations making up a capacitor are recognized widely. Air conditioning tends to minimize difficulties of that nature and at the same time assures maximum stability and reduced maintenance for the precision testing equipment. These factors together with the over-all improvement in the amount of production and quality level of the product more than justify the cost of air conditioning.

Among the advantages resulting from the adoption of the silvered mica construction from the manufacturing viewpoint perhaps the most important was the marked reduction in the number of operations and the time required to effect final capacitance adjustment. This is accomplished readily on silvered mica by scraping off a portion of the electrode area while the unit is connected to a capacitance bridge. Adjusting to close tolerances in the case of clamped mica-foil types was an exceedingly tedious job requiring considerable skill. Another advantage of silvered mica is the fact that stacking to the desired capacity value may be carried out with considerable precision and with no necessity for close thickness limits on the mica. Manufacturing yields were greatly improved, the cost was reduced, and difficulties caused by aging out of limits almost were eliminated by the adoption of silvered mica.

In an article by Doctor A. T. Chapman the manufacture of these capacitors is discussed in more detail.[3]

REFERENCES

1. MICA CONDENSERS AS STANDARDS OF CAPACITY, Harvey L. Curtis. United States Bureau of Standards, Washington 25, D. C., reprint 137, April 22, 1910.

2. MANUFACTURE OF SILVERED MICA CAPACITORS, A. T. Chapman. *Electronics*, McGraw-Hill Publishing Company, New York, N. Y., November 1945.

Electronic Register Control for Multicolor Printing

Discussion and author's closure of paper 46-132 by W. D. Cockrell, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 617–22.

S. L. Burgwin (Westinghouse Electric Corporation, East Pittsburgh, Pa.): I wish to compliment W. D. Cockrell upon the clarity and organization of his paper. I found it unusually easy to read and his method of illustrating the operation of the electronics circuits is very easy to follow.

Early in this paper Cockrell states that the register must be within three thousandths of an inch to be acceptable. We, at Westinghouse, in developing apparatus for similar application in conjunction with the Cline Electric Company have adopted the same value as an objective, although it is my personal opinion that the average reader will not be disturbed unduly until the mis-register exceeds ten thousandths of an inch. I would like to ask Cockrell how he obtained this particular value.

In our early development at Westinghouse we likewise found vibration of the scanning head to be a serious problem and we likewise were forced to shock-mount our amplifier. It is amazing how nearly similar our early designs were to that shown in this paper. However, such shock-mounting resulting in a slight uncertainty or variability of position of our scanner which made it difficult to reduce vibration to a satisfactorily low level and still realize the goal of three thousandths of an inch register. Consequently, we found it advisable in a later design to develop other methods of solving the vibration problem. Perhaps Cockrell has been more fortunate than we have been in this respect.

In our early development we ran into considerable difficulty in maintaining the steep wave front of the voltage pulses generated by the change of light on the phototube, resulting from interelectrode capacity of the first amplifier tube. This resulted in a decreased amplification and a greater sensitivity of register to line voltage changes. Although our circuit differed considerably from that described in this paper, the circuit shown may be subject to the same difficulties.

In Figures 8 and 9 of the paper, voltages produced by web and cylinder scanners are assumed to change instantaneously. Cockrell no doubt has made this assumption to simplify his explanation. I would like to point out that in a more rigorous analysis the actual rates of change would have to be taken into account as they depend upon the width of the slits in front of the phototubes, which are not negligible, and to some extent upon interelectrode capacity of the amplifier tubes, which also may not be negligible. While Cockrell has not stated the widths of these slits I assume that they are in the order of half the width of the register mark. If so, the pulses from the web scanner will be more nearly triangular than square and the error indication would be more nearly proportional to the square of the error than the first power. On this basis the regulator cannot be classified strictly as a proportional regulator but lies between the proportional and dead zone types.

Only a statistical study of the register of a large number of specimens can prove whether the objective of three thousandths of an inch has been attained, and even the results of such a study might not be conclusive as there is always some change in dimensions of the paper from one printing cylinder to another. Whether or not this objective of three thousandths of an inch register has been attained, observation of the operation of the register control described by Cockrell leaves little room for doubt that it is a considerable improvement over manual control.

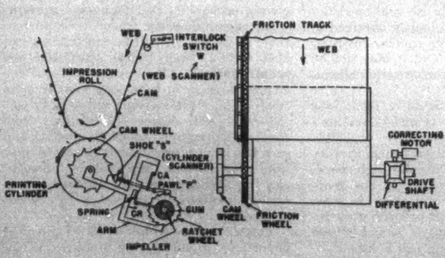


Figure 1. The impression and printing cylinders with imaginary electromechanical equipment for automatic color register

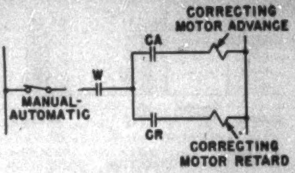


Figure 2. The wiring circuit of the contacts shown in Figure 1

CA.	Cylinder advance contact
CR.	Cylinder retard contact
W.	Web interlock switch

K. P. Grenfell (General Electric Company, Schenectady, N. Y.): It might be of some assistance in understanding the action of the photoelectric color register control, described in Cockrell's paper, to visualize an electromechanical analogy. The problem is to provide a means of keeping the web in a definite longitudinal relationship to the printing cylinder so that the cylinder will superpose its impression exactly over that printed on the web at the previous cylinder.

Suppose it were possible to attach a friction track on the web, as shown in Figure 1 of this discussion, and to have this mesh with a friction wheel on the cylinder shaft. Once the proper relationship were established, register would tend to be locked in, as the web would be geared to the printing cylinder.

In order to get this gearing into phase, and to keep it there whenever it tried to slip, devices would be needed to detect out-of-register conditions. Imagine a series of teeth or cams attached at regular intervals to the edge of the web. These correspond to the web register marks. An interlock switch or web scanner marked *W* has its contacts closed for a brief interval each time a cam passes under its roller.

Picture again a wheel or scanner disk solidly connected to the printing cylinder, with equally spaced cams on its periphery. The number of cams on the wheel equals the number of cams on the web passing per revolution of the printing cylinder. The passage of a cam under shoe *S* momentarily closes contact *CA*, and advances pawl *P* on its ratchet wheel. The ratchet wheel carries a container of gum. When the trailing edge of shoe *S* slips sharply off the trailing edge of the cam, contact *CA* opens, pawl *P* turns the ratchet wheel, and the gum closes contact *CR* by means of an imbedded impeller. Contact *CR* is drawn open again by a spring after a time interval approximately equal to the closed time of contact *CA*. The *CA* contact, pawl, ratchet wheel, and *CR* contact are mounted on an arm which is adjustable about the cylinder axis. The *CA* contact and its operating mechanism correspond to the cylinder scanner phototube and optical system. The ratchet wheel and *CR* contact create the live zone for retarding correction—a function actually performed in the mixing panel.

The three contacts are wired up as shown in Figure 2 of this discussion. The coil causing the correcting motor to advance the

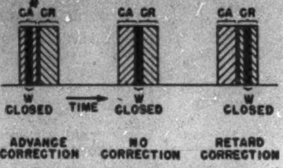


Figure 3. The zones of correction

*Contacts closed

printing cylinder with respect to the web is energized when both *W* and *CA* contacts are closed. The coil for retarding operates when both *W* and *CR* are closed. Under a condition of correct register at the printing cylinder, the arm should be so positioned that shoe *S* drops over the trailing edge of the cam at an instant exactly half way through the closed interval of contact *W*. Reference to Cockrell's paper will show that in the electronic equipment the *trailing edge* of the cylinder disk slits is the indicating point. Under this condition the correcting motor will receive first an impulse to advance, followed immediately by an impulse of equal duration to retard. The net correction will be zero.

Suppose the gearing slips for some reason, and the printing cylinder drops back slightly with respect to the web. Contact *W* now will be closed predominantly during the time that *CA* is closed, and the correcting motor will advance the cylinder till register is restored. Similar reasoning applies to retard corrections. Figure 3 of this discussion illustrates this action. As Cockrell explains, a single correcting pulse lasts only a matter of microseconds. The electronic equipment therefore must integrate these to obtain a signal long enough to operate the correcting motor.

Starting a new run on the press can be visualized easily with this mechanical analogy. With the automatic equipment disconnected by a switch in series with interlock *W*, the operator brings the various colors into register by manual control of the correcting motor. Inspection of the product is his guide. He then moves the arm until shoe *S* slips over the trailing edge of the cam at the same instant that the arm of interlock switch *W* reaches its maximum deflection. Register operation then is switched over to automatic.

W. D. Cockrell: In reply to S. L. Burgwin's question about the 0.003 inch register tolerance which I mentioned, I can say only that it was a compromise, arrived at after discussion of the various types of reproductions to be registered. If the cut consists of fine lines such as the reproduction of an etching or engraving, a difference of even 0.001 inch is noticeable. However, for other reproductions, such as that of a water color having broad areas which blend, 0.010 inch register will be permissible. However, for the usual Kodachrome reproduction, 0.003 inch to 0.005 inch register will do a satisfactory job. The equipment described should hold better than 0.003 inch under constant speed conditions with uniform paper and humidity, but it is believed that the guarantee of 0.005 inch will cover most conditions of constant speed operation encountered in the average color printing plant.

The high sensitivity demanded requires that we shock-mount our electronic equipment. This is shown in Figure 7 in the paper. However, in order to obtain the greatest accuracy of register, the optical system is mounted as rigidly as possible, since it is the position of the optical system (which ends with a slit through which the light is transmitted to the phototube) that determines the accuracy.

Obtaining the proper shape of electric wave front requires careful circuit design and co-ordination with the optical system. However, present day television practice requires the passage of much shorter signals and a much wider range of frequencies. Hence, the circuit requirements, while difficult, are not at all impossible. The projected width of slit is less than one-fourth that of the register mark so that the rounding of the signal resulting from the optical system is not too great a handicap and a reasonably square wave-form results. The control is of the proportional type with a dead zone determined by the low torque required to break the correcting motor starting friction.

K. P. Grenfell's mechanical analogy of the timing and synchronizing principles is quite interesting, and is a reasonably accurate picture of the action taking place. Since most press operators have mechanical rather than electrical training, analogies of this type should be very useful in presenting to them the theory of operation of the register control.

Electrical Accuracy of Selsyn Generator-Control Transformer System

Discussion of paper 46-113 by Harold Chestnut, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 570–6.

C. Concordia (General Electric Company, Schenectady, N. Y.): A paper such as this is valuable first in showing

1. The deviations of the actual component of a control system, the Selsyn, from the ideal proportionality assumed in so many studies.
2. The smallness of these deviations.

It is difficult to obtain in small machines the good wave form ordinarily obtained in large machines by proper winding distribution and pole shaping, and it may not be worth while in view of the deviation magnitudes shown.

Another point that should be mentioned is that it is not possible to make deviations of any magnitude completely negligible simply by gearing the Selsyn system in order to make it run at a sufficiently high speed, since a reasonable accuracy of the gears has to be considered, and there is some optimum division of effort between work devoted to improving Selsyn accuracy and that devoted to improving gear accuracy.

A New Approach to Probability Problems in Electrical Engineering

Discussion and author's closure of paper 46-128 by H. A. Adler and K. W. Miller, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 630–2.

Philip Doane (Consolidated Edison Company of New York, Inc., New York, N.Y.): If we assume infinitely small time intervals so that no two events occur within the same time interval, then origination of a period with r events existing can occur only by an increase or decrease of one event from a state with $(r-1)$ or $(r+1)$ events existing. The numer of periods with r or more events then will be equal to the number with r events existing which originate as increases from the status with $(r-1)$ events existing. For $r>0$ this will be, provided the average length and the frequency of the successive occurrences of any one event are the same for all events,

$$N_r=\frac{n!}{(r-1)!(n-r)!}t^{r-1}(T-t)^{n-r}/T^{n}=\frac{r}{pT}P_r$$

The average length as a fraction of the total time T of periods with r or more events existing is then

$$t_R/T=(P_n+P_{n-1}+\ldots+P_{r+1}+P_r)p/rP_r$$

It should be noted that, as in all of the equations of the paper, the ending of a period with r events existing followed in the next time interval by the beginning of another period with r events existing is considered a division between two different periods.

Equation 12 apparently requires qualification as neglecting those cases in which successive originations of an event occur in time intervals immediately following the end of prior occurrences thereby occupying time intervals otherwise designated as intervals originating periods with the next lower number of events existing. In most practical cases the error would be negligible.

H. A. Adler: In view of several questions and special requests, the following additional background information seems desirable for clarification of the formulas presented in the paper. As an example, if there is a certain probability that four out of n events occur simultaneously, then it is obvious that, in general, it will be much less probable that all four events start and end simultaneously. For events which can change position in time continuously or in small steps, most overlaps will be only partial ones. All degrees of overlaps from small ones to complete ones are possible, and each has its own probability. From all possible overlaps and their probable frequencies the average overlap or average duration of simultaneous occurrences can be calculated.

This average overlap may be of practical importance in engineering problems as can

Table I. Summary of Enumerations for $n = 3$, $t = 4$, $T = 10$

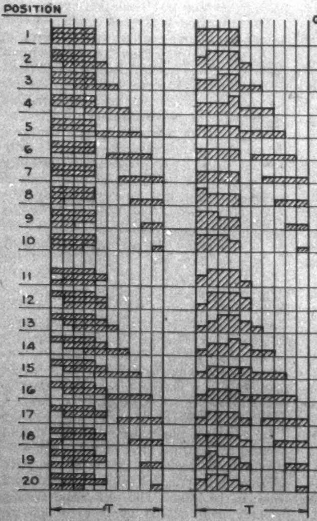
Length of Occurrence in Units	No Occurrence		Single Occurrence		Double Occurrence		Triple Occurrence	
	Number of Times	Total Duration	Number of Times	Total Duration	Number of Times	Total Duration	Number of Times	Total Duration
1	300	300	780	780	660	660	180	180
2	240	480	600	1200	480	960	120	240
3	180	540	420	1260	300	900	60	180
4	120	480	270	1080	90	360	10	40
5	60	300						
6	10	60						
Total	910	2160	2070	4320	1530	2880	370	640

be seen from the following two examples. If a certain probability of simultaneous outages of say four machines on a system were determined, it would make a considerable difference for reserve planning purposes whether the average duration of such an outage were expected to be a matter of a few hours or a few days. Similarly a simultaneous operation of a number of welders on a circuit would have obviously little effect upon voltage flicker if the average overlap were only a small portion of a cycle. Of course for cases where events can change position only in steps equal to their duration, or, in other words, if only complete or no overlaps are possible, then the binomial law gives the complete answer. For these special cases the equations presented in the paper reduce to the binomial formulas. The binomial law is, therefore, a special case of the new equations.

The formulas have been derived empirically by enumeration of a variety of cases. Following is an example to illustrate the method. Consider the case of three events each lasting four consecutive units of time and occurring recurrently once in ten units of time. It is apparent that these three events can take 10^3 or 1,000 different positions in the period of ten units if the events can change position in steps of unit length. Figure 1 shows an analysis of the first 20 of the possible 1,000 positions. On the left side the positions of the three events are shown graphically. The next column shows graphically the consequent amplitude and duration of occurrences. In these graphical illustrations it should be understood that the period T is a recurring measure of time such as hours or years and not a confined length. An event occurring near the end of T then merely would extend into the next period T or it could be considered reappearing at the start of the same period T as it has been done, to save space, in Figure 1. The right side columns of Figure 1 enumerate the duration and amplitude of the occurrences.

The results of the complete analysis of this case are summarized in Table I. It should be noted that the numerals in the last line under the headings "Total Duration" are identical with the numerical value of the numerator of equation 2 for $n = 3$, $t = 4$, $T = 10$, and $r = 0, 1, 2, 3$ while the numerals in the last line in the columns "Number of Occurrences" are identical with the value of the denominator of equation 2 for these cases. Equation 2 has been derived by a systematic analysis of a variety of enumerations such as the one discussed. Analogous methods were used for the duration of the other unit step formulas. The analyses included a number of extreme borderline cases where the formulas might have been expected to break down. However, the formulas were found to be consistent with the enumerations of even such cases.

Figure 1. The first 20 out of 1,000 positions for $n = 3$, $t = 4$, $T = 10$



POSITION	LENGTH IN UNITS NO OCCURRENCE	SINGLE OCCURRENCE	DOUBLE OCCURRENCE	TRIPLE OCCURRENCE
1	6	0	0	4
2	5	1	1	3
3	4	2	2	2
4	3	3	3	1
5	2	4	4	0
6	1+1	4	4	0
7	2	4	4	0
8	3	3	3	1
9	4	2	2	2
10	5	1	1	3
11	5	1	1	3
12	5	1	1	3
13	4	1+1	1+1	2
14	3	2+1	2+1	1
15	2	3+1	3+1	0
16	1	1+1+4	3	0
17	1	1+1+4	3	0
18	2	1+3	3+1	0
19	3	2+1	2+1	1
20	4	1+1	1+1	2

With regard to Doane's comments, the method of statistical equilibrium suggested by him leads to an equation which is identical with equation 5, although this equation has been derived by a different method. He is correct in stating that the ending of a period of r events followed by a period of r events of which at least one is new is considered a division between different periods, see, for example, positions 15, 16, and 18 of Figure 1.

As stated in the paper with regard to equation 12, individual length and order of occurrences of repeated events have no effect on the results. Therefore successive occurrences following immediately the end of a prior occurrence should not influence the accuracy of the results.

4,370,000-Kva Short-Circuit Tests on Grand Coulee 230-Kv Bus

Discussion and authors' closure of paper 46-177 by W. H. Clagett and W. M. Leeds, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 729–34.

T. M. Blakeslee (Water and Power Department, City of Los Angeles, Los Angeles, Calif.): The scheme used to obtain maximum short-circuit current by arranging the test breaker contacts to part within the first half cycle of fault current is a clever method of squeezing out maximum kilovolt-ampere capacity from the available test setup. It is understood that this idea has been used for many years in the Westinghouse high-current laboratory. Nevertheless, if the scheme were known more generally, it would promote a more extensive use of power systems for circuit-breaker testing.

Operating engineers have been somewhat adverse to the use of the power system as a laboratory for short-circuit tests. This aversion is probably justified, because it arises from the idea that there are enough voltage dips in the normal course of events without worrying about deliberately imposed short circuits.

With this anticipated-tripping scheme, breaker tests will produce voltage dips of such short duration that they will not be noticed by the customer. Experience with the 287-kv Boulder circuits, which clear faults within three to four cycles, indicates that these short-duration faults are not

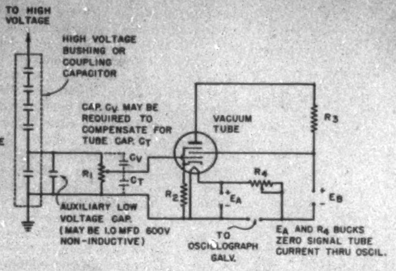

TO HIGH VOLTAGE
HIGH VOLTAGE BUSHING OR COUPLING CAPACITOR
CAP. Cv MAY BE REQUIRED TO COMPENSATE FOR TUBE CAP. CT
VACUUM TUBE
R3
CAPACITIVE VOLTAGE DIVIDER
R1
Cv
CT
R4
R2
EA
EB
AUXILIARY LOW VOLTAGE CAP. (MAY BE 1.0 MFD 600V NON-INDUCTIVE)
TO OSCILLOGRAPH GALV.
EA AND R4 BUCKS ZERO SIGNAL TUBE CURRENT THRU OSCIL.

Figure 1. Bonneville Power Administration oscillograph amplifier basic circuit

noticed. Now, the test engineer using this scheme of Clagett and Leeds can guarantee to the operating force that the deliberately imposed short circuit for circuit-breaker testing will not disturb the customers. This guarantee must be predicated, of course, on the assumption that the breaker test will be successful. The industry should benefit from any scheme that will promote greater use of power systems as laboratories for circuit-breaker testing.

Figure 9 of the paper shows the interruption of 138-ampere charging current at 230-kv. This oscillogram shows the charging current passing continuously through two current zeros following contact parting before interruption takes place at third current zero. Normally breakers interrupt capacitive currents at the first current zero following contact parting, but all breakers, of course, do not maintain this current cutoff without subsequently restriking. This interruption at the first current zero is attributable to the fact that no voltage exists across the breaker contacts at this time as a result of the ability of the capacitor (line or cable) to remain charged at system crest voltage. This interruption at first current zero, even though it may be only momentary, has occurred always in the department's experience with high-voltage charging-current interruption. What is the explanation for the performance exhibited by Figure 9?

Harold L. Levinton (Bonneville Power Administration, Portland, Oreg.): In describing tests of the interruption of short-circuit currents having a magnitude above the present highest rating of 230-kv oil circuit breakers, the authors cover a subject which will be of greater concern as operating companies continue to grow. I concur with the authors regarding the general practicability of constructing 230-kv oil circuit breakers rated above 3,500,000 kva. Also, I believe that in developing breakers suitable for higher ratings, interrupter and operating mechanism improvements will result in even better and more reliable breakers. Operating companies are to be commended for the encouragement they provide designers and manufacturers in co-operative testing which has resulted in more reliableand often in faster breakers than the latest present-day relay systems which often require a longer period to operate than the breakers which they control.

There are two points of general interest which I wish to discuss.

1. The definition of "interrupting time" provided in present standards is not sufficiently complete to cover some types of breakers now manufactured. For example, the 1945 American Standard C37.4, paragraph 4-5.7.4 on "interrupting time" states

A. "The interrupting time of a circuit breaker is the interval existing between the energizing of the trip coil at rated voltage and the interruption of the circuit." A further explanatory note contains the following:

B. "This (interrupting time) is the summation of the opening time and the arcing time."

The "opening time" is defined by the American Standards Association as the interval between the energization of the trip coil and the parting of the arcing contacts, and the "arcing time" as the interval between the parting of the arcing contacts and the extinction of the arc.

The two quotations given above are contradictory where resistors are introduced into the current-carrying circuit during the interruption, in that the extinction of the arc between the arcing contacts (see quote *B*) does not result in interruption of the circuit (see quote *A*); the circuit is not interrupted until the resistor current is broken. My question then is, "Is the interrupting time of a breaker the interval between the energizing of the trip coil and the interruption of the arc in the arcing contacts, or does it extend to the complete interruption of the circuit including any period of resistor current?"

2. From observation of many tests on the Bonneville Power Administration system, I believe that the question of the interruption of line-charging or capacitive current by circuit breakers is about as important as the interruption of high values of fault current because of the possible damage to the breaker (which has been observed more than once) and to connected system equipment, damage that can occur as a result of voltage stresses which can be present as a result of restriking during the interruption of line-charging current. In my opinion, any high-voltage breaker test should include a large number of tests covering the interruption of line-charging current for both short and long lines in order to establish definitely the performance of the test breaker in this important service. Furthermore, I believe that the standards should contain data regarding this important service and that breaker-rating classifications should contain definite minimum requirements with regard to line-charging current interruption.

For use in oil circuit breaker tests and in other tests where appropriate, the Bonneville Power Administration developed a small, inexpensive electronic amplifier device which in connection with the capacitor voltage divider tap of bushings can give an accurate replica of the high voltage on the bushing over a range of frequencies from direct current to 100 kc or more. As normally used, without frequency-correcting capacitors the amplifier is linear from direct currents to well beyond the 10,000-cycle range of high-frequency electromagnetic oscillographs. One of these amplifiers was used in the subject tests to give an oscillographic record of the voltage across the breaker contacts during the interruption of line-charging current. Other uses have included oscillographic recording of the trapped voltage on a transmission line during line-charging current interruption, accurate representation of line and bus voltages during fault conditions, and transient voltages present in static capacitor installations. The device which is made of inexpensive radio components is a simple direct-coupled amplifier. Its basic circuit is shown in Figure 1 of this discussion. Tests have shown it to be considerably superior to capacitor potential devices and to have applications where potential transformers are not suitable. Its cost is less than that of a capacitor potential device. When appropriately used with a suitable electromagnetic oscillograph, it is believed to give results which from a practical standpoint, except for impulse testing, approach those of cathode ray equipment for power system testing.

Philip Sporn and **H. P. St. Clair** (American Gas and Electric Service Corporation, New York, N. Y.): The authors of this paper and the organizations they represent are to be commended for carrying out a courageous job of circuit-breaker testing and thereby making a definite contribution to the circuit-breaker art. The interruption of 10,500 amperes at 230 kv, representing a new high in kilovolt-amperes, seems to offer no difficulty with a modern highly efficient interrupting device. Since substantially higher currents, over 14,000 amperes, have been cleared successfully at 138 kv, and since modern high-voltage interrupters are acknowledged to be essentially current-interrupting devices, there does not appear to be any reason why ratings even higher than 5,000,000 kva at 230 kv cannot be realized. It will be recalled that the need for developing 5,000,000 kva 3-cycle breakers was suggested in a paper,[1] as well as in the closing discussion pertaining to this paper, presented before the Institute on January 1945. In subsequent papers Leeds also indicated his confidence in the feasibility of developing breakers not only for 5,000,000 kva but for possibly as high as 7$^1/_2$-million kva as well.

The situation at Grand Coulee, which necessitated extensive bus sectionalizing to limit the concentration of short-circuit capacity on any one bus section as dictated by limitations of circuit-breaker interrupting capacity, is one that we may be facing increasingly in the future with the inevitable growth of power systems. We agree fully with the authors of this paper that important advantages in operating flexibility will be obtained by having available circuit breakers of sufficient interrupting capacity to avoid the necessity of using more or less complicated bus-sectionalizing schemes. As we have pointed out in other discussions, already we have had occasion to feel the lack of operating flexibility in the sectionalizing scheme we were forced to employ at our Philo plant in order to keep interrupting duties within the maximum of 2,500,000 kva, which at that time was the largest capacity available at 138 kv. We should like to emphasize again that in order to meet future requirements it is none too soon to consider the development of 5,000,000-kva breakers at this lower voltage.

In carrying out these tests at Grand Coulee, the authors have made an interesting deviation from the procedure we have followed in making full-scale circuit-breaker tests in the field. The individual testing of each pole of the breaker by line-to-ground tests undoubtedly insures a greater likelihood within a given number of shots of giving each pole a test somewhere near maximum severity. At the same time, this method may have the advantage that the shock or disturbance from a system standpoint is somewhat less than that involved in 3-phase tests. On the other hand, we have felt that the carrying out of a larger number of full 3-phase tests, with the likelihood that in one or more of these tests each of the three poles would be stressed close to maximum severity, would provide the best over-all series of tests for the breaker. If the total number of tests is to be limited, as appears to have been the case at Grand Coulee, the method adopted of testing each pole individually may be the best procedure.

We have observed also that the tests were confined to the standard duty cycle with a 15-second interval between shots. Without attempting in any way to depreciate the value of these tests it appears to us, in the light of the increasing use and importance of ultra high-speed reclosing, that an even greater contribution to the circuit-breaker art would have been made had it been possible to arrange for rapid reclosing tests at these high short-circuit values. Beginning with the first major field test which we carried out in 1925, we invariably have included at least a series of shots as closely spaced as our control arrangements would permit. While true ultra high-speed intervals were achieved only in later tests, these intervals were always much smaller than 15 seconds. It is our hope, therefore, that if another series of tests of this nature is carried out, it will include either a number of shots typical of the full ultra high-speed reclosing duty cycle or at least a series of tests at much closer intervals than 15 seconds.

Reference

1. Field Tests and Performance of Heavy-Duty High-Speed 138-Kv Circuit Breakers—Oil and Air-Blast, Philip Sporn, Harry P. St. Clair. AIEE Transactions, volume 64, 1945, pages 401–10.

W. H. Clagett and **W. M. Leeds:** Blakeslee refers to Figure 9 of the paper showing the interruption of line-charging current and asks for an explanation of the absence of any attempt to interrupt at the first current zero after contact parting. This is accounted for by the purposely delayed pumping action of the oil flow piston in the interrupter, which begins to function only after the contacts have separated approximately one inch. Because only two interrupting units are used per pole, the voltage gradients across the contact gaps are correspondingly higher than for multi-break designs, so that even a small voltage is sufficient for arc reignition before the oil flow is started. Another factor involved is the small high frequency transient appearing across the breaker contacts at current zero attributable to the shift in bus potential by the amount of the voltage rise through the series inductive reactance of the generators and transformers. In the case in question, this rise measures about 10 per cent of the generated voltage, and, assuming 100 per cent overswing in a 1,000-cycle frequency oscillation, a build-up of 40 kv crest in 500 microseconds would be possible. This could cause an arc reignition within an interval too small to observe on the magnetic oscillogram.

Levinton asks whether the definition of "interrupting time" of a circuit breaker should include specifically the time to interrupt the current through special resistors inserted momentarily in the circuit by the sequence of contact opening. Where the residual resistor current is comparable in magnitude to load or line-charging currents, it would seem logical to include the disconnection of the resistors in the interrupting time. In any case, we agree that it is a matter which should receive attention by the Standards committee for clarification.

As Levinton suggests, it is very desirable to include in any program of field testing of high-voltage circuit breakers a considerable number of line-dropping tests, interrupting the charging current to both short and long lines. The Bureau of Reclamation is well aware of the importance of such testing and has made many series of line-dropping tests on the breakers in the Coulee Dam switchyard. In the absence of a cathode ray oscillograph, the accurate measurement of line and bus transient voltages has been greatly facilitated by the use of the amplifier developed by the Bonneville Power Administration and described in Levinton's discussion.

Sporn and St. Clair seem to favor 3-phase testing over separate tests on individual poles as providing the best over-all check on breaker performance. Where the maximum fault power desired is obtainable as a 3-phase short circuit, this is certainly the case. However, it should be pointed out that the principal reason for including single phase tests at Grand Coulee, in addition to full power 3-phase tests, was the 30 per cent increase in current obtainable on single line-to-ground short circuits as compared with 3-phase faults.

The recommendation that high-speed reclosing duty cycles be included in future high-power field tests of this nature is in line with the trend in high-voltage system operation. While the operating mechanisms on the line breakers in the Coulee Dam switchyard are not designed for high speed reclosing, consideration has been given to revamping them for such duty. In view of current discussions concerning the appropriateness of present de-rating factors for circuit breakers operating on high-speed reclosing duty cycles, the inclusion of such tests wherever possible in field testing programs should be of real value to both operating engineers and circuit breaker designers.

Simplicity in Protective Relaying

Discussion and author's closure of paper 46-173 by Lloyd F. Hunt, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 995–1000.

E. E. George (Ebasco Services Incorporated, New York, N. Y.): This most interesting and valuable paper by L. F. Hunt should be required reading for all relay engineers and all planning engineers.

Without derating the qualifications of Hunt or his associates, it may be suggested that a large part of their success in protection is due as much to their influence on system planning as to the simplicity of their relay schemes.

With due modesty, the discusser may recall his often repeated statement that it is easy to design power systems that cannot be relayed by experts.

Heretofore, relay engineers generally have had too little interest and influence in planning bus layouts and allocating circuit breakers. Hunt's statement that relay engineers should prevent as well as minimize system faults is unanswerable. And what method of prevention is more effective than proper planning of bus and switching facilities? Complicated, expensive, and unreliable relaying is unavoidable if the system plan includes such things as

1. Three-terminal lines.
2. Bus tie or transfer circuit breakers.
3. Transfer or by-pass disconnecting switches.
4. Insufficient line circuit breakers.
5. Slow-operating circuit breakers.

Experience on other transmission systems confirms Hunt's comments on the advantages of

1. Sensitive ground protection.
2. Reduction of backup protection.

Figure 1 of the paper shows several interesting features.

1. The use of ring busses on the newer stations (see also Figures 5 and 6).
2. The use of loop transmission with line circuit breakers at junctions.
3. The use of circuit breakers on lines rather than on equipment.

The suggested use of over-all generator and transformer protection may be questioned by some engineers if it leads to the omission of the generator circuit breaker. The time constant of large generators is sufficient to do considerable damage in case of transformer trouble.

The suggested bell alarm scheme is better than the use of a series bell relay, but not as good as a combination of both—particularly in older stations, where trip coils, batteries, and circuit-breaker mechanisms are none too reliable.

Without forgetting that some of the newer and more complicated relays are essential for certain popular layouts on many systems, it would appear that the Southern California Edison Company is planned and relayed by engineers who know system maintenance and system operation.

W. R. Brownlee (The Commonwealth and Southern Corporation, Jackson, Mich.): The author has presented an interesting review of the relaying systems which have been developed for application to the conditions encountered by his company. It is agreed generally that simplicity in itself is highly desirable, but the so-called simpler devices are capable of performing only relatively simple functions under ideal conditions. For example, the author has avoided the use of distance relays, the inference being that there are no intermediate steps between a plain overcurrent relay and carrier current relay protection. It seems that much better over-all simplicity could be secured by using distance relays, which may be more complicated in theory but not in actual operation, rather than to face the necessity of having station operators transfer the protection of transmission lines from one set of relays to another (as described by the author) whenever system generating conditions change.

Reference is made to sensitive ground protection of distribution circuits. While adequate methods for sensitive protection of 3-wire systems, or 4-wire systems with the neutral grounded at one point only, have been available for many years, the sensitive protection of the almost universal 4-wire multigrounded distribution system is still a most difficult problem, and it is believed that the author's scheme does not apply to these more common systems.

Reference is made to a distribution protection sensitivity of 1-ampere primary with a statement that one of the requirements for this type of protection is for the condition where higher voltage distribution circuits are used jointly with communication circuits. Such an inference is most unfortunate because representatives of the power industry and the telephone industry have agreed that positive protection which will operate on a current of 50 amperes or less at a time of not longer than five seconds is entirely adequate for the higher voltage joint use installations.

A performance record of 99 per cent is mentioned. Is this based on the principles outlined in the report "Uniform Records of Electrical Operation of Power System?"[1] This method of judging performance considers the over-all system and not just the relay. For example, if a relay of the overcurrent type operated precisely on its characteristic to trip a circuit breaker unnecessarily merely because the generating conditions in effect were not those considered when setting the relay, this report would consider the operation incorrect on a system basis.

One of the objections cited by the author to distance relays, or even directional relays, is the uncertain performance of older types of directional elements at low voltage. As early as 1930, directional distance relays were available which would operate positively with all three conductors of a high voltage line tangled together within the first span from the station. The case of a 3-pole gang-operated grounding switch causing practically zero voltage at a station readily has been taken care of elsewhere by providing an auxiliary switch on such a line grounding device to trip the associated circuit breaker without the necessity of any relay action.

Naturally, different fundamental conditions may result in differences in relaying methods. For example, the isokeraunic level of the Southern California Edison system is indicated as between 5 and 10, whereas corresponding values for the midwestern area are from 35 to 55. This and other factors may account for the following items in which our practice differs from the author's:

1. Phase faults must have positive protection.

2. The economical operation of plants, especially where steam power is a heavy proportion of the total, requires that relaying systems be able to meet constantly changing conditions.

3. A radial or simple loop system is relatively easy to relay but may be very costly in the major equipment required to serve a given load.

4. Many applications are found where distance relays are the proper solution, these usually being of the restricted circle variety.

5. The large increase in cost to provide carrier current relaying usually is not justifiable merely to provide simultaneous clearing of both ends of lines which are too long for pilot wire protection, but are justifiable to insure maximum load carrying and swing current ability.

6. Stability studies including a-c calculating board readings of relaying quantities are highly desirable to insure that the relaying system does not become the limiting factor in transmission system emergency capability.

Finally, it is believed that the pendulum has swung somewhat away from having all system planning depend quite so heavily on relaying principles. Many years ago system planning in general paid little attention to relaying or even to emergency operating methods, and experience demonstrated that these must be planned together. Based on progress in the relaying art at that time, rather severe limitations sometimes were imposed on system planning by relaying considerations, and this resulted in uneconomical major expenditures in certain cases. With improvements in relays and methods, and with reasonable co-ordination in planning, it is possible to plan economical power systems in which the relay system will not cause any substantial limitation in emergency system capability.

Reference

1. Uniform Records of Electrical Operation of Power Systems, National Electric Light Association publication 289–52.

Ben C. Hicks (Shawinigan Water and Power Company, Montreal, Quebec, Canada): Agreement no doubt will be quite general regarding the importance, within reason, of simplicity in protective relaying, but a question arises as to what constitutes simplicity and how it best may be attained. In striving for simplicity, sight should not be lost of the fact that the fundamental aim is to secure the most effective over-all protection possible within economic limits. It is doubtful if this end can be achieved by selecting and codifying a few principles of protection, enforcing adherence to these principles alone and disregarding all others. On the contrary, it is probable that the many and often complex problems of system protection can be solved most successfully if approached with an open and untrammelled mind, aware of all the ramifications of the problems that may arise and equally aware of and free to use any of the multitudinous resources that may be available to aid in their solution.

L. F. Hunt's outline of practice on the Southern California Edison Company system presents both similarities and contrasts to practice on the system of the Shawinigan Water and Power Company. Among the similarities may be noted:

1. It is considered to be of vital importance that the control of all system protection be centralized under one authority.

2. It is considered to be impracticable to protect the system completely against all conceivable hazards, and hazards that are likely to have only a minor effect on the security of the system at large often are accepted.

3. Considerable attention is given to the attainment of uniformity in the design and handling of details of protection. An effort also is made however to prevent any attempt at the achievement of uniformity from interfering with effectiveness.

The contrasts are more numerous and include among others, the following:

1. Engineers of the Shawinigan Water and Power Company do not consider it essential that all trip circuits should be locked in.

2. The use of hand reset trip relays is avoided entirely.

3. Little or no justification is found for providing an alarm, actuated by an auxiliary switch, to indicate the opening of circuit breakers.

4. An effort is made to avoid the use of a multiplicity of alarms, which would tend to be confusing to the operators. As far as possible, a common alarm bell is used, often supplemented by some form of annunciator scheme.

5. Considerable importance is attached to the provision of backup protection.

6. Distance relays are used extensively for line and stand-by protections. Current relays alone are not considered adequate for the protection of any important transmission line, except in the case of a few short tie lines which are provided with pilot wires and a differential current protection.

7. Either or both differential and split phase protections are used on synchronous machines.

8. An effort is made to approach simultaneous tripping of both line ends, and tripping in sequence is avoided whenever possible.

Hunt states that approximately 99 per cent perfect relay operation has been obtained over a long period of time. This is a most enviable record. It would be interesting to know the basis of reckoning used in arriving at this figure.

L. F. Hunt: It is gratifying to have such liberal discussion on this paper. E. E. George has read between the lines and understands the intent of this paper and fully appreciates the difficult problems involved.

There is not nearly the discrepancies between the relay protection systems of the Southern California Edison Company and others as brought out by W. R. Brownlee and Ben C. Hicks. If all the references were read a similarity would be appreciated more. As brought out in the paper, the prime purpose of protective relays is to trip circuit breakers to isolate faults. The alarm circuits are not as important. It is agreed that multiplicity of bell alarms is not good; it has been standard setup for 20 years on the Edison system to use only one audible alarm with annunciators for individual indication. The use of hand reset relays individually is not tolerated. As brought out in the paper, only one reset is used for all differential tripping relays in one station. This form is more reliable than the the complication of automatically resetting these relays.

Carrier relaying started on the Edison system in 1926, which is about six to eight years before other applications were used. During this time a carrier system which is very simple was developed by the engineers of the Edison Company. Brownlee is quite correct, relative to the choice of relay systems having no immediate step between plain or balanced current protection and carrier current relay protection. The carrier current relay system developed by the Edison Company has practically the same cost as the so-called distance relay system. The cost of the coupling capacitor can be charged justly to line potential and synchronizing. Therefore it is agreed with Brownlee that carrier current protection is better than the distance relays systems.

The percentage indicated is a very simple result of arithmetic. If all faults are cleared perfectly without any extra switches tripping, the result would be 100 per cent. The longest time at this value was eight consecutive months of operation. However, if 100 faults occur and on one fault the isolation was correct, but an extra switch opened even though no extra load was dropped, the record would be 99 per cent correct. In other words the percentage equals the number of correct fault isolations divided by the number of faults.

As Brownlee states the case, "but the so-called simpler devices are capable of performing only relatively simple functions under ideal conditions," may be true, but with no major system outage in ten years and a record of 99 per cent correct isolation by simple relays seems answer enough.

Transmission Rating of Telephone Systems

Discussion and author's closure of paper 46-163 by W. A. Codd, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 694–8.

Arthur Bessey Smith (Automatic Electric Laboratories, Inc., Chicago, Ill.): In general there are two plans for testing the transmission of a telephone system.

1. Compare an unknown with something similar which is for the time regarded as the standard, at least for the purpose of measurement. The unknown may be judged to be better or worse than the standard of reference.
2. Test the performance of an unknown system without comparing it to anything else and rate it by an independent number or per cent.

The first voice and ear tests followed the first method. The repetition rate, immediate appreciation, and syllabic tests follow the second method as well as some of the machine tests.

In common with others we formerly used only voice and ear tests both for research and development and for factory inspection. But even in those days, mere loudness was never satisfactory to me. For that reason I worked out, adopted, and applied for many years the use of three carefully selected words: three, six, and eight. They were selected because they contain a large number of frequencies which are essential to clear articulation and understanding. These three words are of about equal loudness, a factor of great importance to the listener. They were pronounced carefully and with the same loudness, first on circuit *A* then on circuit *B*. The listener was asked to listen with this aspect in mind: "If I had to receive an important message and get it correctly at once, which circuit would I prefer?" He was directed not to choose the circuit which was the more pleasing to him. The choice of which circuit delivered the message more certainly was made entirely on a utilitarian basis.

Moreover, the voice of the speaker in the tests had to be of the ordinary volume which would be used by the public as far as we at that time could determine it.

For a long time this method gave excellent results, which agreed well with the performance of systems in public use. Though it was a comparison method, it is evident that it duplicated some of the features of the immediate appreciation test described by Hughes (reference 1 of the paper).

For the many who cannot afford the very expensive and time-consuming repetition-rate method, the immediate appreciation method deserves careful study.

W. A. Codd: Doctor Smith has pointed out that the articulation, repetition rate, and immediate appreciation tests give figures of merit which express performance in absolute terms as a rate or percentage for a given set of conditions and therefore is not inherently dependent upon comparison to a standard or reference system. This has some value in that it gives an indication of the degree of approach to perfection in terms of the criterion adopted. It should be noted, however, that in the application of the results of the repetition rate tests and also of the immediate appreciation technique, as proposed in the present paper, a comparison method is used. This is evident from the fact that interpretations of the test data to obtain effective ratings are made in terms of the decibels attenuation of distortionless trunk, which when added to or taken from a given system will make its performance in terms of the given criterion equal to that of a reference system. The significant difference between these tests and the older comparison type tests, such as the voice-ear volume tests, is in the test criterion adopted, namely, some measure of intelligibility instead of volume or loudness.

Dangerous Electric Currents

Discussion and author's closure of paper 46-112 by Charles F. Dalziel, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 579–85.

J. J. Smith (General Electric Company, Schenectady, N. Y.): We are indebted to Professor Dalziel for his contribution in this paper to further knowledge of what constitutes dangerous electric currents. A better picture of this question is being obtained gradually from such studies. Some points still are missing in the spectrum of the effects of the electric currents. As I see it, there are several thresholds in which we are interested.

The first is the threshold of feeling. This is of interest because it is essential that the person using an electric appliance must not get the sensation of a shock. Experimental work has shown that if the current is limited to 0.3 milliampere the danger of such a sensation of shock is small. This limiting value of 0.3 milliampere has been adopted by a joint committee of Edison Electrical Institute, National Electrical Manufacturers Association, and Radio Manufacturers Association and is covered in NEMA publication 105.

The next threshold is the "let-go" threshold. This is of interest because below this level the person experiencing the shock is able to let go freely, although he may experience a very distinct sensation of shock. Thus, if the current did not exceed this value, a person might feel more free in handling it, than is the case in which there is a possibility of higher currents. This threshold has been given by Professor Dalziel as approximately nine milliamperes in men and seven milliamperes in women.

The next threshold in which we are interested is not quite so clear. This is a result of the lack of good data in the region between 10 milliamperes and 100 milliamperes. In his tests, Professor Dalziel subjected some subjects to shocks up to 25 milliamperes with no ill effects when they were released from the current, which was on only a short time. I have heard of cases where people were subjected to 50 milliamperes with no lasting effects. It would be desirable to have definite data to show whether shock in any part of the region between 10 milliamperes and 100 milliamperes is liable to cause cessation of breathing so that artificial respiration must be applied. Some of those with whom I have discussed this question were of the opinion that such currents applied momentarily could stop respiration, others felt it could not. Whether this does or does not occur in this region is of importance in connection with men who normally work on electric circuits under suitable precautions but because of the nature of their work, they are liable to come in contact with these circuits once in a while. They normally would have another man working with them who can remove the power quickly should they be unable to let go.

The next threshold is that of ventricular fibrillation which has been estimated by Ferris and Williams at about 100 milliamperes. In the present paper Professor Dalziel extends our knowledge of this to include the effect of the duration of the shock. I understand that ventricular fibrillation does not occur often when the current through the body is greater than about one ampere but would like Professor Dalziel's comment on this point.

The next threshold is that for which fibrillation does not occur but where the current is large enough to render the victim unconscious, so that artificial respiration must be applied. I wonder if Professor Dalziel has any figures he could suggest on this subject.

The next is the threshold for burning. It would seem desirable to have an estimate of the amount of current which when passed through the body for even a short time might be expected to cause burns.

Because the use of 60 cycles is so common, it would be well to have all these thresholds first established for this frequency. At a later date similar thresholds might be given for direct current and various other frequencies.

With the development of radar during the war, it would be desirable to have similar thresholds in connection with such pulsed currents.

L. P. Ferris (New York, N. Y.): Professor Dalziel's paper is based practically wholly on the data presented in the 1936 paper by Williams, King, Spence, and myself (reference 4 of the paper). No new experimental data are presented on this subject, Professor Dalziel confining himself to a reanalysis of our data and the presentation of the results in the form of mathematical equations. No single conclusion of the 1936 basic paper is invalidated or altered by Professor Dalziel's reanalysis. Despite this gratifying fact, I must take exception to some applications of Professor Dalziel's methods to our data. It undoubtedly would be convenient if the results could be expressed in simple mathematical form. Professor Dalziel has done this, but I question his justification, particularly when applying it to the data on effect of shock duration on threshold current, Figure 14 of the 1936 paper. The authors of the earlier paper were hopeful of discovering some simple correlation between minimum fibrillating current and shock duration. We could not, however, draw a straight line on log-log paper which we thought would correctly represent the experimental data presented by our Figure 14. However, Professor Dalziel seems to have done this in his Figure 6. He throws out in his analysis the experimental data which we designated as mode *A*. I cannot take much exception to this because inspection of our Figure 14 shows that we gave but little weight to the mode *A* data in drawing our curve. Professor Dalziel also chooses to omit certain other experimental points, namely our 0.47-second shock tests on sheep. Whether or not this is justified, we should have further experimental evidence before the effect of duration is represented as a simple exponential function.

It should be remembered that in changing the duration of shock from several seconds to a fraction of a second, entirely different conditions are encountered. The several-second shocks blanket one or more complete heart cycles, whereas shocks of a small fraction of a second duration are only effective in causing fibrillation when they occur during the partial refractory phase of the heart cycle, which is about 20 per cent of the whole. In passing from one of these conditions to the other, it should not be surprising if the curve of fibrillating current versus shock duration changes slope.

It is to be hoped that Professor Dalziel and other investigators may be able to carry further the experimental investigation of electric shock. Further data on the effect of wave form and frequency of shock current, as well as shock duration, would be desirable. Also desirable would be a study of resuscitation in cases of known ventricular fibrillation with and without electric counter-shocks. Progress with measuring instruments and in the electronic art in the last ten years should make it possible to control and measure shocks with greater precision than was possible before 1936.

Charles F. Dalziel: J. J. Smith's definitions and discussion of the significance of the various current thresholds are exceedingly valuable. His discussions are stimulating because the answers to many of his questions are unknown. With the exception of one paper (reference 4 of this paper) very little data are available for currents in excess of about 25 milliamperes.

In 1936 letters were sent to several neon sign manufacturers in the attempt to determine the hazard from 30-milliampere high reactance 5- to 15-kv sign transformers. No fatal accidents were reported. L. T. Jones, a manufacturer, reported as follows:

> From personal experience in the industry I believe it is safe to assume that in each plant, each day, some one or more persons receives a shock from a 30-milliampere transformer. Insofar as I have been able to determine, not one person has been injured . . and I have been able to find no record of a fatality.

In preparation of this closure I communicated with major sign manufacturers and others in the San Francisco area who are cognizant of accidents in this field. Although accidents involving burns were mentioned, no one could recall a single fatality on 30-milliampere equipment.

Observations during the let-go experiments indicated that currents in excess of about 20 milliamperes from hand to hand were very painful and the accompanying muscular contractions made breathing difficult if not impossible during the shock; however respiration resumed automatically upon interruption of the current.

Determination of the threshold currents required to produce respiratory inhibition or unconsciousness would be a difficult but valuable undertaking. The threshold current producing burns is probably very low, as feeble sparks, if continued for more than a few moments, cause burns. Dr. W. B. Buchanan, Hydroelectric Power Commission of Ontario, reports that under suitable conditions 0.35 milliampere will cause some carbonization of wooden insulator pins, but with 0.7 milliampere the action was much more definite. Reference 4 and Figure 16 of the paper and the accompanying discussion is offered in reply to the question regarding the decrease in likelihood of inducing ventricular fibrillation at currents considerably in excess of the threshold.

Perhaps the best reply to L. P. Ferris is to state that in the beginning of many engineering analyses approximations often are tolerated pending completion of conclusive research. Certainly conclusions based on limited data are open to question, and especially where human life may be at stake. Future investigations may indicate that fibrillating currents cannot be expressed accurately by simple mathematical relations or as straight lines. In the meantime, a tentative answer may be better than no answer, especially if it is based on a conservative analysis of all available experimental data. Attention was called to the uncertainties of the analysis in the paper, and I should like to repeat that application to man is entirely conjectural. However, only one out of the 99 experimental points in Figure 6 of this paper falls below the one-half per cent line for the 57.4-kilogram sheep. With this single exception the proposed fibrillating current thresholds, as indicated by the one-half per cent lines of Figures 5 and 6 of the paper, include a reasonable factor of safety. It would be a simple matter to draw other lines having a greater factor of safety, if other investigators thought this advisable.

The same data were used in the preparation of Figure 14 by Ferris, King, Spence, and Williams, as in Figure 6 by the author. The dissimilarity of the figures is because of the method of plotting the data, and although log-log scales were used in both, the presentations are sufficiently different to preclude direct comparison. In Figure 14 the minimum current causing fibrillation was plotted versus the ratio of the duration of the shock to the length of the heart beat. Throughout my paper I took the fibrillating current equal to 1/2 (minimum current causing fibrillation plus the maximum current not causing fibrillation). In Figure 6 fibrillating currents were plotted versus shock duration in seconds. Hence the differences in appearance of the two figures and resulting curves do not represent as wide a divergence of opinion as might be concluded from casual inspection.

Although the original data designated as mode *A* and the 0.47-second shock tests on sheep were not used in either the statistical analysis or in determining the deviation curve of Figure 3, for the reasons discussed in the paper, these data were included in Figure 6. Without exception all the data furnished the author by L. P. Ferris and associates were included in Figures 5 and 6. All data were considered when deciding just where to draw the best line to represent the general trend of the responses.

It might be well to mention that the purpose of evaluating fibrillating current thresholds is to establish (if even only approximate) the most dangerous current threshold to permit a more accurate analysis of accidents, and to add to the general fund of knowledge on electric shock. Considering the various factors and their uncertainties, the author still believes that the proposed analysis may be used without excessive error to estimate the probable fibrillating hazards for large groups. The analysis should not be used to calculate an individual's chance of survival or to excuse a careless workman. Because of the limited information on fibrillating currents, and the lack of any data on the currents causing respiratory inhibition and unconsciousness, no human should be exposed purposely to electric currents approaching fibrillating thresholds.

The author appreciates the many suggestions and comments given this research, and would welcome receipt of factual information regarding serious accidents. This invitation is made with the hope that eventually sufficient quantitative data may be assembled to permit arriving at more accurate answers to some of these questions.

Carrier Supervisory Control of Pumping Station Over Power Cable

Discussion and author's closure of paper 46-174 by W. A. Derr, W. A. Keller, and H. A. Hedke, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 699–705.

Abe Tilles (Pacific Gas and Electric Company, San Francisco, Cal.): The establishment of a satisfactory carrier channel over power cable, as here reported, is a development of great interest. It arouses further concern because some earlier attempts in this direction gave very unfortunate results.

In making use of a line-to-ground channel the path provided by the power conductor is excellent. The return path, except where there is an uninterrupted lead sheath, is quite a different matter. For a conductor not enclosed in steel pipe or ferrous armor, general ground paths are likely to be available. This is particularly true for underwater cable.

For cable without continuous lead sheath and enclosed in ferrous armour or pipe, the return path is comprised of several parallel paths none of which are favorable. These include

1. The high reactance paths outside of the iron.
2. A thin path on the inner surface of the iron.
3. A thin path spiraling around the cable along any shielding tapes.
4. A path lengthwise down the cable along shielding tapes which overlap or spiral in contrary directions and contact in a basket weave effect.

If a path of type 4 of relatively low impedance is presented, the 60-cycle losses may be increased to an undesirable degree.

In view of the pioneer nature of this installation it would be very useful if the authors, or the cable suppliers, in discussion could add to this paper by giving in detail the physical characteristics of the cable. What kind and thickness of lead sheath, shielding tape, and so forth, are each of the two pieces? What are the test or calculated data showing where the return current of the carrier channel goes? How were the losses calculated in design?

W. A. Derr: The construction of the two sections of cables used on this installation is as follows:

1. *Armored submarine cable* (22,800 feet in length), three conductor, American wire gauge 4, 19 strands of 0.0469-inch copper, $^{12}/_{64}$ inch Buna *S* insulation. Each conductor taped with a rubber-filled cloth cable tape, conductors cabled with jute fillers, double wrap of cable tape, layer of saturated jute. Birmingham wire gauge 8 galvanized-steel wire armor.
2. *Jacketed cable* (12,200 feet in length), three conductor, American wire gauge 4, 19 strands of 0.0469-inch copper, $^{12}/_{64}$ inch wall of Buna *S* insulation. Each conductor taped with cable tape, cabled with jute fillers, layer of cable tape, $^{8}/_{64}$-inch rubber sheath over-all.

The construction of both cables was determined mainly for 60-cycle power operation. Prior to manufacture of the cable, the phase-to-phase attenuation at carrier frequencies was calculated from the following formula:

$$\text{Decibel loss per loop mile} \approx 4.34\ \sqrt{C} \times \sqrt{\frac{R^2}{L}\left(1+\frac{\overline{PF}^2}{2}\right)+\omega^2 L\overline{PF}^2+2\omega R\overline{PF}}$$

where

C = mutual capacity in farads per mile
R = resistance in ohms per loop mile
L = inductance in henries per loop mile
$\overline{PF}$ = power factor
$\omega = 2\pi$ times frequency in cycles

In any calculation of attenuation the main difficulty is in establishing proper constants. Table I gives the choice of cable constants used in calculating cable losses for both of these cable sections.

The calculated losses were low enough to assume that satisfactory carrier operation could be effected from phase to ground despite additional loss in the ground return circuit. No attempt was made in calculations or tests to determine the actual path of the carrier currents in the ground circuit.

Our experience with cables has indicated that discontinuities in metallic sheaths result in very high attenuation as mentioned by Tilles. However, where there is a continuous sheath, whether lead or iron, the attenuation is not excessive.

It is expected that the carrier ground current in the submarine cable portion of this installation flows along the sheath with relatively small loss. The carrier ground current in the jacketed cable portion of this installation probably flows some distance from the cable, the actual path being determined by the reactance and resistance involved. The loss on this installation in the ground return circuit for the jacketed cable is not excessive since the over-all attenuation is not exceptionally high.

Table I. Cable Constants Used in Calculating Cable Losses

	60 Kc	100 Kc	150 Kc
Resistance (ohms/loop mile)	11.6	15.7	17.9
Inductance (henries/loop mile)	0.00123	0.00123	0.00123
Mutual capacity (μf/mile)	0.148	0.148	0.148
Power factor	0.010	0.010	0.010

Note: Method and values for calculated losses were obtained from the Okonite Company, Passaic, N. J.

Fault Location and Relay Performance Analysis by Automatic Oscillographs

Discussion and author's closure of paper 46-109 by H. P. Dupuis and W. E. Jacobs, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 442–6.

G. D. Floyd (Hydro-Electric Power Commission of Ontario, Toronto, Canada): Automatic oscillographs have been used on the 230-kv network of the Hydro-Electric Power Commission of Ontario for about 15 years to provide a record of system performance including relay behavior and to locate faults. At the present time there are four such oscillographs installed at three of the six terminal stations of the network. Four additional oscillographs are on order. The average length of the trunk lines forming this network is about 200 miles.

Our experience has been much the same as that of the authors in that we found when preparing sets of fault current–distance curves from network analyzer data that mutual induction between circuits had a marked effect on the current distribution. One of our difficulties always has been to provide for the wide difference in current to be measured on a faulty circuit from the near, and far ends, and on parallel and sound circuits. The ratios may be in the order of 40 to 1. Consequently, the deflections obtained for the smaller of the two currents may be very small with a relatively large error in their measurement. It is this source of error, which is inherent, that may limit the application of the method of ratios, suggested by the authors, in our case. This method is sound where it can be applied and overcomes the error caused by fault and tower footing resistance. We agree that this source of error is usually present. Our method of decreasing this error has been to use residual currents on tie lines other than the circuit involved in the fault and to record total transformer bank ground current at two of the stations where oscillographs are located. The indicated locations of fault from all these readings are averaged. The number of faults located during the three years that this method of fault location has been given a serious trial is much better than expected considering the possible sources of error.

The usual rapid clearance of faults on this 230-kv network gives only one or two cycles of record before the first circuit breaker clears. The rapid clearance permits us to use the initial values of current, obtained from the analyzer, without modification. Figure 8 of the paper shows that currents were read after 19 cycles of fault, at which time some decrement in current would be expected. A further advantage in the use of the ratio of currents is indicated, although not claimed by the authors, in that the error introduced in the direct measurement by decrement is eliminated. Our experience has been that a great deal of valuable data may be obtained by use of these instruments at the cost, however, of considerable effort in their application and maintenance.

W. E. Jacobs: The operating department is interested in oscillographs mainly from the operating standpoint. That is, they are interested so far as the information obtained will make it easier in locating faults and analyzing system disturbances. The confidence which the operating men have in the oscillograph reports has been growing steadily with the improvements in the methods of interpreting the oscillograms.

It is the practice of the Consumers Power Company to make an inspection of a line as soon as practicable after a fault. This inspection usually will be made before the oscillograms can be developed and analyzed. If no evidence of the fault is located on this inspection, then another inspection of a short section is made after the location is calculated from the oscillogram. This second inspection covers a section two to three miles each side of the calculated location and is made very carefully, using a high power glass and at times climbing towers. Usually this inspection is made on a bright day. In case of a single line-to-ground fault, the field man can be told the position of the faulted conductor. The primary purpose of this second inspection is to find hidden damage and to determine whether there is some condition that can be corrected to prevent a recurrence of the fault.

We have found it very convenient when specifying the fault locations to use tower numbers as well as the miles from the terminals and have made-up charts showing tower numbers by miles. When the damage is found, the report lists the tower number which gives an accurate check on the calculated location. Since many tower lines run diagonally across country, this designation of fault location by miles alone may be subject to considerable error.

The servicing and calibrating of the oscillograph is handled by relay maintenance men. However, certain other men are instructed in taking the film out of the oscillograph and placing it in special mailing containers. These men are members of the plant organization where the instrument is located and are readily available to send the film for developing immediately after a storm. These men remove and mail the film after each storm or disturbance or when specially requested to do so by the dispatcher. Occasionally one storm will operate the oscillograph more than once. Each instrument is equipped with a station identification letter and a clock stamp which are both very necessary for identifying films after they are developed. A log of the oscillograph operation is kept for each instrument and supplements the time stamp. A signal bell is connected to sound when the oscillograph operates.

The films are developed at our laboratory at Jackson, but since one of the instruments is nearly 150 miles from Jackson it may require two or three days before films are received and developed. Consideration has been given to having the films developed locally and telephoning the measured value to the dispatcher to speed up the fault analysis. However, the number of permanent faults has been low and in most cases of permanent fault other lines are available to maintain service. Therefore the need for this method has not been great. We would welcome discussion on how other companies arrange this service and also on the methods used to develop the films. Our present method of developing the films, especially the longer rolls, is not entirely satisfactory from the standpoint of the time required.

As evidence of the speed of clearing faults, exactly 50 per cent of the 256 faults analyzed in the past five years caused so little damage that they could not be located by the initial field inspection. Part of those faults which are not located on the initial inspection are located at the end of the lightning season when a pole-top inspection is made. In the past two years a factor equal to kiloamperes squared times cycles has been calculated for each fault. We have termed this "damage factor". No definite conclusions have been drawn, but a broad general statement can be made that if the damage factor is below 50 there probably will be no visible damage, and if it is above 200 the damage probably will be found. Further experience may show that these limits can be closer. This is a burning factor and at the present time we have no data to correlate it with insulator breakage.

Of the 256 faults 96 per cent involved ground current. Only four per cent of the faults were phase to phase. Lightning caused 80 per cent of the faults; outside interference such as birds and machinery, caused nine per cent; and wind accounted for 7.4 per cent. Approximately four tripouts were caused by lightning per 100 miles of line per year for this 5-year period, and approximately five tripouts resulted from all causes.

In the operating department we feel that the greatest value of the oscillograph other than its value in locating faults is the information that is obtained when a fault is not cleared properly. In most cases these data definitely point to the cause of incorrect operation or failure to operate so that immediate steps can be taken to prevent their recurrence.

High Voltage D-C Testing of Rubber-Insulated Wire

Discussion of paper 46-121 by W. N. Eddy and W. D. Fenn, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 576–8.

C. T. Hatcher (Consolidated Edison Company of New York, Inc., New York, N. Y.): The authors have presented data on d-c testing of rubber-insulated cables at the factory in order to eliminate inherent defects in the insulation. This method appears to be a step in the right direction as any test method which does not affect the insulating qualities of the rubber insulation, but still acts as a means for eliminating defects, is quite desirable.

C. E. Betzer (Commonwealth Edison Company, Chicago, Ill.): This paper gives some valuable information on the suitability and effects of voltage tests for factory process and acceptance testing. It indicates that such tests might leave undetected some serious imperfections in the insulation. Since the operating stresses for rubber-insulated cable are very low and breakdowns in service are caused mainly by oxidation (either by natural aging or by corona) and not by electrical puncture (except by lightning), these serious imperfections have not attracted attention in service. However, as the ability to withstand oxidation is improved, there should be an increasing necessity to eliminate these inperfections and then to reduce insulation thicknesses.

The Insulated Power Cable Engineers Association specifications for ozone-resistant type of insulation require an a-c acceptance test for cable rated at five kv and higher of about 90 volts per mil, or a d-c test of about 270 volts per mil after 24 hours immersion in water. The authors show that insulation with a fine needle puncture might withstand this test and with a U-knife cut easily could pass the test if squeezed in the manner common on a reel containing several turns of wire. Furthermore, it is shown that insulation with air pockets leaving only ten per cent of sound rubber withstood tests at 340 volts per mil direct current after 12 days of immersion, which indicates that it would have had a very large margin over the required test after one day of immersion. In another case insulation with an iron chip through two-thirds of the thickness withstood three tests at 570 volts per mil direct current during 12 days immersion.

Defects in rubber insulation almost never are found after the wire is shipped from the factory. The operating stresses are less than five volts per mil for low voltage wire or cable, 12 to 15 volts per mil for 5-kv cable operated at four kv, 22 to 25 volts per mil for 15-kv cable operated at 12 kv. Failures caused by electrical puncture by operating voltage (other than oxidation) almost never occur. When they do occur the cause seldom is found. Carbonized paths cannot be seen in the black insulation, and the insulation cannot be opened to view in the same manner as paper or varnished cambric insulation.

This paper suggests that some upward revisions of the acceptance test voltages for rubber insulation would be helpful.

The Electrical Performance of Ceramic Dielectrics at Elevated Temperatures

Discussion and authors' closure of paper 46-117 by H. A. Frey and J. M. Jesatko, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 911–20.

C. M. Sherer and **D. I. Smith** (Pennsylvania Water and Power Company, Holtwood, Pa.): One of the practical accomplishments of this investigation was to provide a satisfactory solution to a problem of porcelain failures where porcelain was used as an insulator across high voltage in the hot stack gases of electrostatic precipitators.

Drive shafts made of glazed porcelain were used between the electrode rapping device and driving motor in the Cottrell precipitators installed at the Holtwood steam plant. The shafts were 54 inches long, and had a 5-inch outside diameter with a $1^1/_{32}$-inch hole through the center. Operating voltage across the shafts was nominally 70,000 volts direct current, and operating temperature was of the order of 500 degrees Fahrenheit. Average life of the shafts was $2^1/_2$ to three months, and when failure occurred it was usually in the form of a clean shear break or in the nature of shattering. The breakages at first were thought to be caused by mechanical stresses, and it was hoped to reduce them by shortening the shafts. Whether or not a shorter shaft would be dielectrically satisfactory with respect to flashover was to be determined by wiring strips of sheet

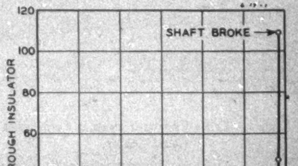


Figure 1. Dielectric failure of porcelain shaft in Holtwood precipitator 3

Boiler started at 3:17 a.m., April 23, 1945 metal to the sides in such a manner that half the length of the shaft would be short-circuited. Two shafts so treated and placed in service were found broken three days later, the exact time of failure not being determined. A third one is known to have failed sometime during the first day, and still another seven hours after being placed in service. Fusible temperature pellets which had been placed on two of these shafts showed that the porcelain under the metal strips had been at a temperature of between 500 and 550 degrees Fahrenheit, which was the same as the ambient gas temperature. The bare portion of the shafts, however, had reached a temperature of between 650 and 700 degrees Fahrenheit, thus indicating a dielectric failure because of runaway as found by Frey and Jesatko in their experiments. This half in each instance shattered into many small fragments in contrast with the shielded half which remained intact.

It now appeared that the failures were produced electrically rather than mechanically, and the cause was assumed to be a thermal runaway effect initiated by conduction current through the porcelain at the elevated operating temperature. As proof of this an idle shaft with half its length similarly short-circuited was placed in the hot gases without mechanical load and the normal operating potential of 70,000 volts applied across it. A recording milliammeter was connected in series with the shaft on the ground side to obtain a graphic record of the conduction current. The bare half of the shaft shattered $12^{1}/_{2}$ hours after the boiler was fired. The curve in Figure 1 shows the progressive increase in conduction current for the last $5^{1}/_{2}$ hours preceding the failure of the shaft. During this interval of time current increased from 1.33 to 109 milliamperes, or 8,000 per cent, with most of the increase occurring in the final minutes before failure. The runaway effect thus is shown clearly.

An experimental shaft made of zircon then was installed in regular service in place of a porcelain shaft. It was somewhat different in design from the porcelain shafts, being $4^{5}/_{8}$ inches square with a $3^{1}/_{2}$-inch hole through it. When it was installed, half of its length was shorted out in order to duplicate the conditions under which the porcelain was tested. Under the same operating conditions as the porcelain, this shaft lasted 28 days or 672 hours as compared with ten hours for porcelain (average of two for which the time of failure is definitely known). On the strength of this, all the porcelain shafts were replaced with shafts of zircon. Three of these now have been in service for periods of $7^{1}/_{2}$, 13, and 14 months, and to the present time there have been no failures.

Our operating experience therefore supports the findings of Frey and Jesatko. Zircon has shown its superiority over porcelain under the temperatures and potentials encountered in the Holtwood precipitators. The authors are to be commended for having established the characteristics of ceramic dielectrics so that their performance can now be predicted at elevated temperatures and operating voltages.

Julius J. Torok (Corning Glass Works, Corning, N. Y.): This is a very welcome paper as there is very little information available on the electrical performance of various ceramic bodies at elevated temperatures. The authors are to be complimented on their analysis of their test results.

Table I

Temperature, Deg F	Resistivity, Megohms
400	5.1×10^{7}
600	2.0×10^{5}
800	4.1×10^{3}

Their data on glass checks roughly those obtained at the Corning Glass Works laboratories. The deviations are probably caused by the lower voltage gradients used in the Corning tests. Tests at Corning have been made on all glasses used by the electrical industries for such applications as radio tubes and lamps. Since in many of these applications the glasses are subject to high temperature uses, it was necessary to make such tests as well as develop new glasses which would give satisfactory results under a wide range of requirements.

One of the borosilicate glasses which has good electrical characteristics at elevated temperatures has resistivity characteristics as given in Table I.

It can be seen that the resistances of this glass is much higher than that of the general purpose borosilicate glass tested by the authors.

An analysis of the data presented by Frey and Jesatko on glass indicates that the borosilicate glass which shattered at the instant of failure was a tempered piece of glass. Highly tempered glass will shatter if a deep penetration into the volume of the glass occurs. If the authors still have some of the pieces of that glass it would be desirable if they would send them to Corning for stress analysis.

Paul M. Ross (The Ohio Brass Company, Barberton, Ohio): The authors have performed a meritorious job in contributing basic data on the temperature–resistance performance of porcelain. Figure 3 presents the voltage gradient versus resistivity characteristics of wet process electrical porcelain. In the normal operating temperature region for porcelain insulators, 150 degrees Fahrenheit and lower, the resistivity of the porcelain varies markedly with change in voltage gradient. Will the authors explain this characteristic?

Several of the specimens were carried to electrical failure at elevated temperatures. Was the type of failure obtained similar to that found on service insulator failures which involved attachment parts and external surface leakage conditions in addition to internal porcelain conduction?

While a discussion of the oil puncture test, as described in American Standard C29.1–1944, paragraph 4.7, is not contained as a part of this paper, do the authors find any evidence from their temperature versus voltage tests that explains puncture locations outside of the enclosed head section of suspension insulators as obtained on the oil puncture test? Do the authors' tests indicate that such puncture locations may be indicative of the electrical quality of the porcelain?

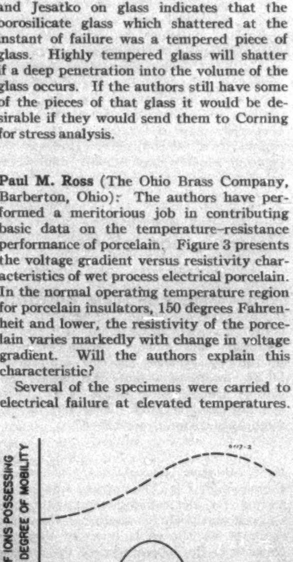


Figure 2. Distribution of ion mobility at normal and elevated temperatures

H. A. Frey and J. M. Jesatko: The authors wish to thank the discussers for their interest and for the pertinent facts brought out in their discussions.

C. M. Sherer has shown the practical application of these data to the solution of a baffling problem involving the failure of precipitator shafts operating at high temperature and high voltage. The curve of leakage current versus time presented by Sherer shows very clearly the characteristic of a long time runaway. The point of upward break in this curve corresponds to the point at which some element in the dielectric has exceeded the critical temperature T_c. The physical breakage resulted from the forces of differential expansion.

Paul M. Ross has asked several pertinent questions. The first of these requests an explanation of the variation of resistivity with voltage gradient. Any such explanation must be hypothetical. We know that the situation exists but at present we only can theorize as to its fundamental cause. As Doctor Torok has pointed out in his discussion, the conductivity of these dielectrics is believed to be caused by an ion drift. Let us suppose that there are in the dielectric body ions of varying degrees of mobility. If the relative mobility of the ions is distributed at random it would be expected that the distribution would follow a normal distribution curve as shown in Figure 2. Ions possessing a high degree of mobility would be influenced by very small voltage gradients while those having lower degrees of mobility would require higher voltage gradient.

Also as the temperature is raised the mobility of all ions is increased and some of those which are fixed relatively firmly at low temperature succeed in breaking loose from these bonds resulting in a shift in the distribution curve as indicated by the broken line in Figure 2. If these suppositions are made it will be seen that a resistivity voltage gradient curve will be obtained for any temperature having the typical form illustrated in Figure 3 of the paper.

Ross also asks whether the failures obtained at high temperature resemble those found on service insulator failures. Some of the tests described in the paper were made on actual service insulators of the type to which Ross refers. The failures on such insulators were exactly like low temperature punctures in such insulators. The point of failure was marked by a fused path in the dielectric sometimes accompanied by radial cracks. The location of the punctures in these cases were always inside the metal enclosed head. On some long insulators, such as typical standoff insulators used in radio applications, the failure more nearly resembled a mechanical break and is caused by differential expansion. These failures were much like those described by Sherer.

The authors are inclined to believe that the punctures in oil outside of the enclosed head to which Ross refers are caused by either dielectric defects or localized stresses set up by the presence of the oil.

Parallel Circuits in Servomechanisms

Discussion and author's closure of paper 46-107 by H. Tyler Marcy, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 521–9.

D. G. Soergel (nonmember; Washington University, St. Louis, Mo.): This paper is a very complete study of the solution of servo design problems through the use of inverse transfer functions. In particular, Marcy has been very general in his approach and has assumed a general feed-back transfer function $K_fG_f(j\omega)$ which is characteristic of many industrial applications, for example, automatic temperature control where the characteristic periods of the controlled element and the measuring feed-back element are often of different values. The analysis is complete in that the author has derived and plotted inverse system functions to indicate static stability and has also shown how a knowledge of the quantities α, ζ, and μ_n is necessary to estimate the system transient responses. The results are plotted in nondimensional form, making them applicable to any problem.

I believe that nondimensionalized system transfer charts plotted for basic lead controllers, integral controllers, combined integral and lead controllers and tachometer derivative stabilized controllers with $\beta = -1$ would be of much help in assisting designers in the design of controllers for a variety of servomechanism applications. Marcy's paper has illustrated the approach to use to develop such charts and opens the way to a systematized procedure to follow for rapid controller design. Problems involving different feed-back transfer functions would become a special design problem, but experience in application may show in the future which types (such as the process illustrative example of Marcy's) would merit and would lend themselves to chart design.

All in all, Marcy's paper is a definite contribution to the engineering phase of servo design, and a utilization of his approach to the problem should result in data which will fit very well into high production design methods. The analysis used is of course characterized by a great deal of mathematics, and it should not be required that production engineers spend time developing the basic equations used. Rather, they are more concerned with the results to be expected of a particular system and whether or not the results are compatible with the particular application they have in mind. This information should be available in chart form.

It is agreed generally that since all physical systems are nonlinear systems, the design criterion arrived at through a mathematical analysis based on linear parameters is at the best a close approximation. The important thing, however, is a knowledge of the system trends caused by a change in one of the system parameters. This information is generally available either in the mathematical expressions describing the system or may be obtained from the nondimensionalized charts applicable to the particular system.

H. Tyler Marcy: I should like to thank D. G. Soergel for his interesting discussion of this paper. It is particularly interesting that he believes that the charts can be of engineering use in the design of industrial controllers. The availability of such charts certainly would be more convincing than a recommendation of the normally optimum relationships of circuit time constants. The charts tell at a glance how critical the controller adjustments are.

A word of caution is advisable, however, in recommending the use of charts. Only when the number of variable parameters represented is small can the relationship between them be presented adequately on a single sheet of paper. Thus, control processes that respond to sudden changes of the controller in an oscillatory manner are difficult to treat in a concise way because they can be described only by a statement of sensitivity, undamped natural frequency (or resonant frequency, if easier), and ratio of actual to critical damping (or degree of resonance, if easier). For this type of process a simple proportional controller, aided perhaps by tachometric feedback or lead networks, may not represent the usual complexity of the controller. However, the problem already has become difficult to treat with the aid of charts.

Determination of Cable Temperatures by Means of Reduced-Scale Models

Discussion and authors' closure of paper 46-103 by Andrew Gemant and Joseph Sticher, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section pages 475–82.

R. W. Atkinson (General Cable Corporation, Bayonne, N. J.): The reduced-scale-model testing for determining temperature by Gemant and Sticher is another of the clever methods for the cable industry that has come from the laboratories of the Detroit Edison Company. There is little doubt that, like the earlier contributions, this one will be the source of valuable information to the industry. The reduced time constant accompanying the reduced dimensions is one of the very important advantages that might have been overlooked in a casual consideration.

I have been glad to learn that we may expect to have further data from the setup at Detroit. My further remarks will relate to suggestions toward such future work.

The important data that can be learned from a system such as these authors have developed are not concerned with what happens in the cable itself or in the immediate vicinity but rather with what happens in the surroundings—the duct structure or the earth around a buried cable. The performance of the cable and its immediate surroundings, specifically its performance within the duct, readily can be superposed on the data showing rise of duct structure. I have been urging this since 1913 and thus may be prejudiced somewhat, but the plan is so logical that it has come to be generally followed.

To study the rise of the underground structure as contrasted with that of the cable above its immediate surroundings it is necessary to provide test conditions where a larger part of the total rise is that of the structure. In general, this means making tests where many loaded cables, instead of only one, are placed in a duct bank. It is pertinent also that it is usually only in duct banks where several cables are installed that significant current limitations are imposed by the heating of underground cable. Where a cable is not heated by adjacent cables, the underground portion will operate, during most of the year, at a much lower temperature than the end portions.

It also will be very desirable when making these determinations on the model setup to arrange a comparison with a heavily loaded bank of commercial cables. While the theoretical basis for correlation of the model tests with commercial sized installations seems sound and it is perhaps adequate, the authors themselves have recognized the desirability of having an experimental verification of this correlation. The experimental comparison shown in the paper does not give a basis for indicating that a good correlation will exist between model tests and the performance of a loaded commercial duct bank.

It has been of interest to note that the indicated rise of sheath temperature over adjacent duct wall is considerably lower than I have indicated or than generally has been accepted. This is in large part because most of the tests reported cover ducts purposely made wet, but data are shown that suggest that the accepted constants should be subject to review even for dry ducts. However, it will be well for the authors to consider the desirability of giving greater attention to making tests in which there is no free water in the ducts. It can be argued that many ducts or that most ducts are wet. However, in most cities I believe that the serious problems of current limitation occur in ducts that are fairly dry, or in others when they are dry though they often may be wet.

I have gone into detail in these suggestions because I believe the proposed method is suitable for a practical solution of a wide variety of problems relating to the current rating of cables. In a way it may be considered as a mathematical instrument or an integrator set up for solving our special cable thermal problems. It serves this purpose, however, which is not accomplished by the purely mathematical calculators. It does, by its physical setup, what is required of a mathematical physicist in setting up the mathematical equations to be solved by the mechanical calculating machines.

Herman Halperin (Commonwealth Edison Company, Chicago, Ill.): Anyone who has to do frequent work on current rating of cables and is faced with the uncertainties of some thermal constants will welcome such research as covered in the paper by Gemant and Sticher.

As pointed out by the authors, reduction in the dimensions of model installations is permissible accurately only where conduction alone is involved, while for thermal circuits including convection and radiation the conditions are more complicated. For example, the International Power Cable Engineers Association standard on current ratings provides that the thermal surface resistivity of lead sheath shall be expressed as follows:

> "A straight line from 650 at cable diameter equal to zero inch to 1,200 at cable diameter equal to 1.75 inches. For diameters larger than 1.75 inches, a flat value of 1,200 degrees centigrade per watt per square centimeter is used."

It is apparent, therefore, that a reduction in size of a model would give erroneous results. Similar limitations apply to cables in pipes or three cables in a tube, and other problems of this nature for which information is important.

The size of the cable is also important in electrical problems where proximity and skin effects are a factor, for example, for three cables installed in an iron pipe.

The authors have recognized that corrections need to be made in adapting results of tests with models to field conditions, and I am mentioning these points just as a matter of emphasis.

With regard to the use of a box 15 feet long, five feet wide, and four feet high, filled with soil, we wonder whether such a model actually represents the current flow under field conditions. Especially during the summer months, the level of the ground water may be an important factor affecting heat dissipation.

It would be of interest if test results with these models could be compared with the usual formulas used in the industry so that, if necessary, corrections could be made in the constants used in this formula.

For a given conduit cross section and occupancy by cables, it has been found in many field surveys that the heating constants will vary considerably. This variation is a function of the type of soil, the percentage of moisture in it and other things. In Chicago we have found that the variation might be three to one or even more, excluding conduits that are submerged. It therefore has seemed necessary to make field surveys of conduit temperatures, particularly during the summer months, in order to avoid using excessively high heating constants of the conduits in determining load ratings.

C. T. Hatcher (Consolidated Edison Company of New York, Inc., New York, N. Y.): The determination of cable temperatures in duct banks is a problem which all operating engineers have to determine. System investments are affected materially by proper determination of cable temperatures because it is quite important not to install more copper than is required and, conversely, it is important that sufficient copper is used in order that the life of cables is not shortened. Any method which will help in determining proper temperatures is quite desirable.

The authors have outlined a very interesting development and it provides a relatively quick and inexpensive method of obtaining almost exact thermal limits for particular cables in specified buried duct structures under various load cycles. The data may be used for transient as well as steady state determinations. The method does not appear to be flexible enough to permit its use economically for determining loading limits on wide varieties of duct configurations such as are met in station cable designs.

The proposed method apparently can be used for determining short time ratings of important feeder installations where the limiting structure can be determined by observation. It would be quite desirable to be able to extend this method to the determination of cable ratings in complicated station duct structures where little data are now available. The authors have presented considerable data but the form in which it has been assembled requires considerable study to obtain logical conclusions.

L. F. Roehmann (Anaconda Wire and Cable Company, Hastings-on-Hudson, N. Y.): The aim of the authors' method of reduced-scale-model tests is to determine, versus time, the temperature of cables installed in a duct bank under transient or steady state conditions. According to the authors, a reduction of the linear cross-sectional dimensions by a factor α (α smaller than unity) and a reduction of the load currents by the same factor α produces a time contraction by α^2 but leaves the temperatures unchanged. These relations are valid if the thermal parameters of the model are the same as for the full-size installation. They disregard convection and radiation in the duct bank, and magnetic and dielectric losses in the cables.

There are cases in which magnetic and, particularly, dielectric losses may not be disregarded without introducing serious errors. This limitation might be avoided by loading the model cables so that their loss in watts per foot is equal to the loss of the real cables.

Regarding the amount of experimental equipment involved there seems to be little saving as compared with tests on life-size cables installed in a life-size duct bank. Both cases require sources of adjustable current (as many sources as there are cables), thermocouples, recorders, and so forth. A laboratory duct bank has to be cast in a laboratory soil box and laboratory-size cables linearly reduced to about one-fifth natural size have to be manufactured specially.

It would seem to us that either full-scale installations should be tested or, if models are required because of space and time limitations the scale reduction should be greater than one-fifth. If there is a real interest in sheath temperature–time curves, in addition to conductor temperature–time curves, the duct bank should be set up with the cables in natural size thereby avoiding the errors inheren tin a model reduction. However, it may be sufficient to obtain at somewhat reduced accuracy conductor temperature–time curves for transient and steady-state conditions. In that case a model of much smaller dimensions will be adequate. The cables can be represented by heater wires, and the loads can be controlled with small laboratory-size equipment, for example, variable-primary transformers.

R. J. Wiseman (The Okonite Company, Passaic, N. J.): I have had the good fortune to be acquainted with this research since its beginning, discussing the theoretical aspects and later analyzing its test results. I think Gemant and Sticher have introduced a method for studying the heat characteristics of cable systems, both buried and in ducts, which can go a long way to help us solve many of the problems which have confronted us for years.

Discussing first the buried cable case, we have the well-known Kennelly log 4 L/D formula for the steady state case. Some of us think that perhaps a cylindrical log formula will apply equally as well. Also the formula for the transient temperature of a cable before arriving at a steady state as given in the authors' paper is complicated and not easily understood except by mathematical experts. One cannot visualize easily that which is taking place. The authors have shown that a buried cable will follow the exponential integral formula. A few years ago I was interested particularly in the transient temperatures of a buried cable system such as the Bennett oilostatic system. I approached the problem from a cylindrical field viewpoint and had tests conducted by Doctor Paschkis of Columbia University on his thermal bridge. I was able to set up empirical formulas based on the cylindrical log formula which reproduced the actual curves. Later I took the exponential integral formula and also reproduced the curves obtained by Doctor Paschkis. Now I feel that I safely can use either method for determining transient temperatures of a buried cable, but my empirical method is easier to calculate and also gives a better picture of that which takes place. I am hoping, as soon as I have time, to prepare a paper on our investigation.

As the authors point out, it is impossible to express by means of a formula the case for cables installed in ducts. However, the error from the assumption that a cable model will give the desired information empirically is not great as shown by the results if one is willing to accept a difference of three to four degrees centigrade between actual field results and laboratory tests. This error is partially caused by some errors in assumption, for example, the radiation from the surface of a cable and a cable model are not exactly comparable because the emissivity of the surface is a function of the diameter.

The study of the influence of blowing air through a duct or allowing water to flow through is very helpful in proving their desirability where conditions sometimes warrant heavily overloading cables, or where hot spots develop and a quick method must be found to prevent excessive cable temperatures.

The authors are interested primarily in what cables will do under overload conditions, that is, how long a cable will operate at a given load before reaching a limiting temperature or what load is permissible for various times before reaching a limiting temperature. I believe the authors have given us a means in a laboratory for checking the duct heating constants which now are used. At the present time we use values which are not related to the kind of soil or its moisture content; yet for a direct buried cable we want to take into consideration all the variables. This is logical for buried cable systems; why isn't it equally logical for duct systems? By setting up model duct systems with varying numbers of ductways we can make load-temperature tests and then estimate the heating constants. By varying the kind of soil and moisture content we can determine the influence of both.

I wish to congratulate the authors on their unique method for determining the transient temperature of cables without resorting to complicated calculations and a means for investigation of other temperature problems pertaining to power cable systems.

Andrew Gemant and **Joseph Sticher:** We appreciate Atkinson's constructive comments. At present, additional tests on our model setup are under way that meet some of the requirements which he stresses in his comments. In particular, various tests are being carried out on a heavily loaded duct bank which contains eight 3-conductor cables. Part of these tests also involve comparison of model and field data, insofar as these latter are available. Other tests are carried out with one of the ducts dry and the other ducts wet in order to reveal loading limitations in dry ducts as compared with wet ones.

Halperin correctly points out that certain factors, such as skin effect and radiation, require the use of correction factors when determining field temperatures from model data. We recently developed the correction term to be used in reference to radiation, and this term is being used, for instance, in our present tests involving dry ducts. We further believe that our test box is sufficiently large to represent field conditions, since the temperatures in the field are not affected beyond a 12-foot distance from the cable (2.5 feet in the model). However, a larger box could be used easily, if wanted, and water simulating ground water conditions also could be introduced in the model. Halperin's suggestion concerning the use of the model for a redetermination of present-day duct constants will be commented upon in our answer to Doctor Wiseman.

Hatcher emphasizes that our method is not flexible enough to take care of complicated station duct structures. If he has in mind open troughs in which the majority of the heat is lost through radiation, then his statement is correct. For all cases in which radiation inside a duct does not carry more than, say, 40 per cent of the total heat, the model in combination with a correction term can be used.

Roehmann finds only little saving in a model equipment as compared with a field test. It is difficult to agree with this opinion, considering the amount of information now being assembled from model tests which needed about two weeks to carry out, tests which would require a whole year's work in the field. Roehmann further states that conductor temperature data could be obtained from a model much smaller than ours, utilizing heating wires for cables. This statement probably is based on a misunderstanding since theory clearly shows that conductor temperatures can be obtained only if the cable itself is reproduced accurately on a scale reduction basis.

Doctor Wiseman in his comments mentions an investigation of his own which is concerned with establishing an empirical formula for evaluating transient temperatures. We are looking forward to the opportunity of discussing Doctor Wiseman's method more in detail, when his paper will be published. He also emphasizes the possibility of using the model method for checking existing duct heating constants, a suggestion made also by Halperin. We welcome this suggestion as the model method is certainly well-suited for such type of work. We think that it would need careful planning and also a general agreement among various interested companies as to the advisability of such research before steps to carry it out are undertaken.

Dynamic Braking Control of D-C Series Motors—Calculation of Stability Limit

Discussion and authors' closure of paper 46-130 by G. W. Heumann and W. P. Smith, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 454–7.

J. D. Leitch (Electric Controller and Manufacturing Company, Cleveland, Ohio): The authors are to be complimented on a very fine paper containing data of great importance to the designer of d-c crane hoist control. They have chosen to express their results in the per unit system which makes their curves useful over a wide range of motor frame sizes. By an interesting choice of units they have been able to express the torque per ampere of armature current, and the counter electromotive force per rpm for various values of field current in a single motor saturation curve. This should reduce greatly the time required in calculating the speed–torque characteristics of any hoist circuit using a d-c series motor.

The authors point out that they have chosen the one-half hour rating of the motors in defining their base units. This is in accordance with the practice on general purpose cranes and it is important that the motor rating be defined carefully in determining the stability limits of a hoist controller when lowering a heavy overhauling load. In the case of more severely worked cranes, such as bucket cranes, it is customary to choose the motor sizes on the basis of their one-hour ratings while on standby cranes, such as are used in a power station, the one-quarter-hour ratings sometimes are employed. Obviously, in designing controllers for those cranes on which the motors have been chosen on the basis of the one-quarter-hour ratings, greater care must be exercised in assuring that the stability limits are not exceeded.

Using the published characteristic curves for a wide range in sizes of mill type motors we have endeavored to plot the K_s and K_t curves on a per unit basis and found considerable variation between motors of different frame sizes. Though there appeared to be a definite trend as we considered motors of larger sizes, this trend was contradicted in a number of cases. We acknowledge however that we were compelled to perform a certain amount of extrapolation of the characteristic curves and that the accuracy of the method could at best be approximate only. In the light of the authors' work with a wide range in motor sizes and with actual test data at their disposal, would they care to give us some information regarding the variation in the K curves (Figures 2 and 4 of the paper) for the range of motors investigated?

The authors show very clearly the importance of a series resistor R_s (Figure 7) as a means of increasing the stability limit. Would they care to discuss the value, if any, from a stability point of view, to retaining a small amount of resistance in the armature circuit?

G. W. Heumann and **W. P. Smith:** D-c series motors as used on cranes are generally mill or crane type motors for which manufacturers' handbooks publish the ratings not only on a one-half-hour basis, but also on a one-hour and one-quarter-hour basis. It should be noted that although different time ratings may be assigned to a motor, its speed–torque curves and stability limit in terms of rpm and pound-feet are not altered thereby. A given set of motor test

Table I

For Curve Shown in Figure 2 of the Paper

Per Unit Field Amperes	Maximum K/Minimum K
0.30	1.24
0.50	1.11
0.80	1.03
1.00	1.00
1.20	1.03

Table II

For Curves Shown in Figure 4 of the Paper

Per Unit Armature Amp	Max K/Min K 1.00 PU Field Amp	0.5 PU Field Amp	0.3 PU Field Amp
Generating			
2.0	1.2	1.35	1.93
1.5	1.19	1.17	1.38
1.0	1.15	1.26	1.25
0.5	1.11	1.25	1.30
0.0	1.07	1.20	1.29
Motoring			
0.5	1.04	1.10	1.20
1.0	1.00	1.08	1.30
1.5	1.03	1.20	1.60
2.0	1.07	1.28	1.64

data which are carried out far enough to permit calculation of the limit of stability may be expressed in per unit to any base. The per unit values of either test data or subsequently calculated data may be changed to any other base by the relation:

$$PU_2 = PU_1 \times \frac{B_1}{B_2}$$

where

PU_2 = per unit value of given quantity on base quantity, B_2

PU_1 = per unit value of given quantity on base quantity, B_1

It was hoped, when work was first started on this paper, that it would be possible to arrive at average per unit curves which would be sufficiently accurate to apply to a large range of motor sizes. It was found, however, that the curves of individual motors are so different as to preclude this possibility. It also was found that, while there were some definite trends as larger motor sizes were considered, the curves, especially the K curves of the type shown in Figure 4, were of many different shapes. The resulting variations in K value were often so large as to be totally misleading. To illustrate the magnitude of these variations the data shown in Table I and Table II of this closure are presented.

As might be expected, variations are greater for weak field strengths than for heavy field strengths. The general trend with motor size of saturation curves, such as Figure 2, is for larger motors to have lower than average K at field strengths less than 1.0 PU, and higher than average K at field strengths above 1.0 PU. Smaller motors exhibit the opposite tendency Exceptions, however, are notable and frequent. No definite trend with motor size is noted with curves of the type shown in Figure 4.

Resistance in series with the armature only has no effect on the steady-state stability limit. According to equation 11 of the paper, torque is solely a function of armature current. Resistance in series with the armature determines the electromotive force according to equations 12 and 13, and consequently the speed according to equation 10. Therefore, if a circuit were used similar to Figure 6, except with resistance in series with the armature, the maximum torque would not change, but it would occur at a higher speed.

Adding resistance in series with the armature increases the slope of the speed-torque curve so that, with a given no-load speed and a given stability limit, a higher full-load speed is obtained. When switching from the last point to preceding points during deceleration, the presence of resistance in series with the armature has a beneficial effect in that transient current peaks are reduced; therefore it can be said that armature resistance increases the transient stability of the motor during deceleration.

Another advantage is gained by the use of resistance in series with the armature when two motors are connected to a large hoist, forming a duplex drive. Armature resistance will cause the two motors to share their load more evenly than without armature resistance. On a duplex drive the combined stability limit is given by the motor carrying the higher load, and for this reason it may be stated that resistance in series with the armature has the tendency of increasing the stability limit of a duplex controller.

Circuit Cushioning of Gas-Filled Grid-Controlled Rectifiers

Discussion of paper 46-146 by D. V. Edwards and E. K. Smith, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 640–3.

Roman Pogorzelski (Power Equipment Company, Detroit, Mich.): An experience with one of our rectifiers in the field demonstrates very significantly the effectiveness of circuit cushioning as presented in this paper.

The rectifier under discussion uses xenon-gas-filled grid-controlled tubes operated much of the time at approximately 30 per cent of their rating and approaching the initial inverse voltage conditions given in test 6 of Table II of the paper. Without a cushion circuit the tube life did not exceed 500 hours in any case. After the cushion circuit, suggested by D. V. Edwards, was added, the rectifier has operated well over 11,000 hours to the present date with no tube failures. The cushion circuit consisted of an 0.6-microfarad capacitor in series with a 300-ohm resistor across each tube.

With regard to measurements to determine the values of constants in the cushion circuit, it was found expedient to use a method that permits the calculation of these values. Outlined briefly, the method is as follows. The maximum current commutation rate is calculated from the maximum commutating voltage obtainable in the plate transformer and the leakage inductance as measured by a short-circuit test. The value of A_{10} is then computed, and the maximum safe rate of rise of initial inverse voltage is determined for the given tube type. The values of resistance and capacitance are derived from consideration of the transient response of the series R, L, C circuit formed by the cushion circuit and leakage inductance to the suddenly applied electromotive force at the end of the commutation.

J. W. Picking (Reliance Electric and Engineering Company, Cleveland, Ohio): We have designed our standard electronic variable voltage d-c motor drive controller to incorporate xenon-filled grid-controlled rectifiers as manufactured by the authors' company.

Our original life tests and actual sample installations indicated we could expect an average normal life in excess of 5,000 hours on 6.4-ampere *C6J* grid-controlled rectifiers. These tests were based on single-phase full-wave rectifiers supplying power to a d-c motor armature. In all cases the load was such as not to require continuous current conduction throughout the normal operating range.

One of the sample installations was used to drive the work spindle of a precision universal grinder. After operating for several months it was noticed that chatter marks were set up on the ground surface especially on large diameter work. The cause of these marks was traced to angular vibration set up in the driving motor by the discontinuous pulses of rectified current.

To eliminate this trouble sufficient inductance was added in the form of a reactor in series with the motor armature so that current in the circuit was continuous for all values of load and speed. Within one month, approximately 450 hours of operation, both tubes had failed. Tube tests indicated high arc drop. Results were similar on a second and third set of replacement tubes.

D. V. Edwards was contacted and cushioning as covered in this paper was suggested. This was tried and this original set of tubes still are operating, having passed the 8,000-hour point recently.

On the basis of this and similar results on other installations it has become the standard practice of our company to apply cushioning circuits to all grid-controlled rectifiers, either gas- or mercury-filled where continuous conduction is expected at any point in the operating range.

This practice has caused us to redesign all of our anode transformers, adjusting the leakage reactance. Because the amount of cushioning required is a function of transformer leakage reactance we have found that a satisfactory compromise between voltage regulation and required cushioning can be obtained so that the maximum cushioning per tube will be less than 1.0 microfarad in series with less than 150 ohms for all sizes of rectifiers up to 25 amperes at 230-volts output.

Considerable difficulty was experienced in obtaining oscillographic readings for determining the current ten microseconds before cessation of current and the rate of rise of initial inverse voltage. Using a pulse driven sweep oscilloscope of the Dumont 248 type has reduced the taking of these readings to a matter of minutes and is recommended to those interested in studying these circuits in detail. As a warning, distributed capacity of the leads and instrument must be neutralized carefully as the wave fronts are very steep.

D. V. Edwards and E. K. Smith are to be complimented on developing a practical solution to a very difficult problem.

The Frequency Response of Automatic Control Systems

Discussion and author's closure of paper 46-114 by Herbert Harris, Jr., presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 539–46.

D. G. Soergel (nonmember; Washington University, St. Louis, Mo.): Harris has presented a very clear description of a servomechanism analysis technique which very easily can be applied to engineering practice.

It would be an agreeable situation if all information needed to evaluate transient responses could be obtained from steady state frequency response data. Harris' statement that the resonant frequency persisting after a transient condition could be determined from a transfer loci plot and that an increase of frequency response would result in a more rapidly damped system seems very reasonable. However, specific information necessary for the synthesis of a controller such as the system attenuation constant α, the actual time necessary to reach a steady state condition, and the magnitude of initial overshoot would require a transient analysis of the system. Unfortunately, as was stated in the paper, the analytical determination of transient data often involves more work than is practical. The usual procedure then is to design the system from a stability approach and make laboratory corrections for faulty transient operation. If laboratory setups are physically impossible, the only recourse is a mathematical analysis.

A design criterion of 1.5 for the value of M is a good choice and can be determined most easily from a comparison of constant M and inverse loop transfer plots. Harris discussed and illustrated inverse loop transfer functions to some extent but his paper largely was concerned with the determinaand calculation of direct loop transfer functions and the determination of system responses from them. We believe that advantages realized from the use of inverse functions justify their use over direct functions for the reasons given by the author and also for the simple fact that concentric M circles and straight-line constant phase angles are plotted more easily.

A standard procedure to follow is to derive the inverse system transfer function and then separate the constant gain factor from the frequency variant portion. The product then is plotted and the approach of the locus to the M equals 1.5 constant M circle is noticed. If the locus is not tangent, the gain factor may be increased accordingly until tangency is reached, and the intersection of the system locus and constant M circles and constant phase angle lines will determine the system characteristics.

The treatment of inherent controller delays and compensating or antihunt networks is also of interest since it indicates to the designer an approximate order of antihunt values, and also the correct direction for a contemplated circuit change.

In conclusion, the paper does not present anything noticeably different in the treatment of servo design and seems to treat the most convenient approach to the problem, inverse system functions, merely as another way to solve the problem. Actually inverse functions should rate first attention. The paper, is written however, in clear engineering language and discusses quite ably the physical significance of the mathematics needed to calculate loop transfer functions. This is a very welcome bridge in the gap between the high mathematics needed to solve servo problems and the actual values and parameters used to approach physically the ideal expressed by mathematical quantities.

Herbert Harris, Jr.: Soergel points out the difficulties involved in trying to correlate the frequency response parameters such as resonant peak amplitude and frequency with transient response parameters such as attenuation constant and response time. This is a recognized difficulty but it loses much of its importance as the servomechanism designer becomes experienced in comparing the two solutions. With this experience, it is just as easy to think in terms of the frequency response as it is in terms of the transient response. The saving in calculation time is certainly worth considerable effort in achieving a new point of view.

It is true, with some exceptions, that the inverse transfer functions are easier to use than the direct transfer functions presented in the paper. Historically, however, they came into general use several years after my original presentation of this method in a National Defense Research Council report dated December 1941, and consequently there exists much experimental and design data in the form of direct transfer functions. H. T. Marcy has described the use of inverse transfer functions in more detail.[1]

REFERENCE

1. PARALLEL CIRCUITS IN SERVOMECHANISMS, H. Tyler Marcy. AIEE TRANSACTIONS, volume 65, 1946, August–September section, pages 521–9.

The Influence of the Concentration and Mobility of Ions on Dielectric Loss of Insulating Oils

Discussion of paper 46-110 by Bun Po Kang presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 403–07.

Andrew Gemant (The Detroit Edison Company, Detroit, Mich.): Bun Po Kang is to be commended for having obtained a reasonable amount of information on the ionic conductivity of oils, notwithstanding the fact that he had only a 60-cycle power factor bridge and a viscometer at his disposal. His unit of poise-watts which is used as a measure of the ion content of an oil clearly indicates the inherent limitation of power factor measurements. By this means one obtains only the product of ion content times mobility, and a separation of the two in order to obtain absolute values for both is not possible without additional information. This is the reason why the discusser has been for the past few years engaged in a research of introducing electrochemical electromotive-force measurements into the study of insulating oils. It is to be hoped that by this means, just as in aqueous solutions, more complete information on the concentration and mobility of ions may be obtained.

Attention also should be called to the fact that the phenomenon discussed in connection with Figure 6 has been familiar to those who have investigated mixtures of insulating liquids (see Association of Edison Illuminating Companies minutes for 1934, pages 475–8).

Dimensionless Analysis of Servomechanisms by Electrical Analogy

Discussion and author's closure of paper 46-150 by S. W. Herwald and G. D. McCann, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 636–9.

W. M. Pickslay (Allis-Chalmers Manufacturing Company, Milwaukee, Wis.): This paper presents an interesting new tool for the analysis of regulating or closed cycle systems. Most of the work in this new type of analysis has occurred with angular position servomechanisms of comparatively small powers. This is natural because the development of the analysis coincided with the war demand for such particular servomechanisms.

Previous papers of the AIEE including one by men from the same manufacturer have shown that the performance of larger generators subjected to sudden load changes is substantially independent of the voltage regulator. The main factors affecting regulation are the design characteristics of the generator and up to a certain speed, the exciter speed of response. This being the case, I would like to ask if the use of this new tool has resulted in any significant improvement of performance of voltage regulating systems through alteration of the regulator or servomechanism parts or circuits.

G. D. McCann: W. M. Pickslay's question can be answered best by stating that a very comprehensive study of voltage regulating systems has been made recently with the transients analyzer. These data are being presented in a future paper.

Tennessee Valley Authority Hydroelectric Stations—Electrical and Mechanical Design

Discussion and author's closure of paper 46-108 by R. A. Hopkins and H. J. Petersen, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 920–31.

S. H. Mortensen (Allis-Chalmers Manufacturing Company, Milwaukee, Wis.): A brief discussion of the design of the Chickamauga generators shown in Figure 3 of Hopkins' and Petersen's paper may be of interest to AIEE membership. The generators shown in cross section in Figure 1 of this discussion are duplicates rated 30,000 kva, 90 per cent power factor, three phase, 60 cycles, 13,800 volts, 75 rpm, class *B* insulated and guaranteed to carry rated load with temperature rise not exceeding 60 degrees centigrade over 40 degrees centigrade air.

Their design is of the overhung or umbrella type with the thrust and single guide bearings supported on a bracket below the generator rotor. This construction, suitable for large diameter slow speed machines with narrow cores, facilitates inspections and repairs as it permits generator rotor removal without disturbing shaft and bearing alignment, as well as dismantling of the bearing housing shafts and bearings in one piece without changing the unit alignment. It also reduces generator head room requirements to a minimum. With the thrust and guide bearings both mounted in a single water cooled oil bath, no provisions for oil circulation are required. The flat adjustable 8-segment Kingsbury thrust bearing 93 inches in diameter supports the combined weight of the unit's rotating parts as well as its hydraulic thrust, which combined amounts to 2,050,000 pounds. The 85-inch diameter by 14-inch long single guide bearing is made in sections free to pivot and bearing upon a smooth finished surface of the thrust block. The lower portion of the bearing is submerged in the oil bath, and the rotation of the shaft establishes an oil flow through longitudinally spiraled bearing oil grooves adequate for bearing lubrication. The construction of the oil container in conjunction with that of the sectionalized cooling coils and guide and thrust bearings permits removal of the latter through normally covered openings in the container sides.

For shipping reasons the rotors were built with fabricated arms bolted to a shaft-mounted hub. Flanges on the arms support the punched floating type rotor rim that carries the field poles and coils.

The dovetailed field poles have copper damper bars silver-soldered into the pole end plates. The latter are fitted tightly into the spider rim and the two combined completes the winding circuit.

A break rim with replaceable segments is supported on the spider arms.

For shipping reasons the generator stator was made in four sections which were stacked at the factory and wound at the destination. Shims provided between yoke joints provide means for adjustment of core pressure at splits.

The generator ventilating system is of the enclosed circulating type and designed for CO_2 fire protection. Two rotor mounted fans circulate 75,000 cubic feet of ventilating air per minute over the coil and core surfaces through the yoke and eight fin-type water-cooled heat exchangers which absorb the heat losses. From the coolers the air is discharged into an air housing which surrounds the stator and from there it returns to the rotor fans. The air housing which also encloses generator leads and piping allows space for inspections. Access is gained by means of a number of doors sufficiently tight to hold CO_2 when admitted to the ventilating system for fire protection.

The direct driven 300-kw main and 28-kw pilot exciters mounted on a separate bridge are ventilated independently of the main generator by means of separate fans as shown in Figure 1 of this discussion.

A few dimensions and weights of the Chickamauga generators may be of interest. The stator bore is 380 inches; the outside diameter of the air housing is 557 inches; the net stator weight is 194,000 pounds; the net rotor weight with alternator shaft is 579,500 pounds; the air housing weighs 82,000 pounds. These weights combined with those of bearings, brackets, exciters and sole plates bring the total machine weight up to 1,200,000 pounds. The rotor $WR^2 = 75{,}000{,}000$ pound-feet-squared, and it is designed for a runaway speed of 207 rpm.

The characteristics of the Chickamauga generators and exciters which were built some years ago conform to the recent Tennessee Valley Authority specifications given in Hopkins' and Petersen's paper except for the construction of the thrust block, which is a separate casting, and the mounting of the rotor on the generator shaft. The difference between this design and that corresponding to the recent Tennessee Valley Authority specifications becomes apparent by comparing Figures 1 and 2 of this dis-

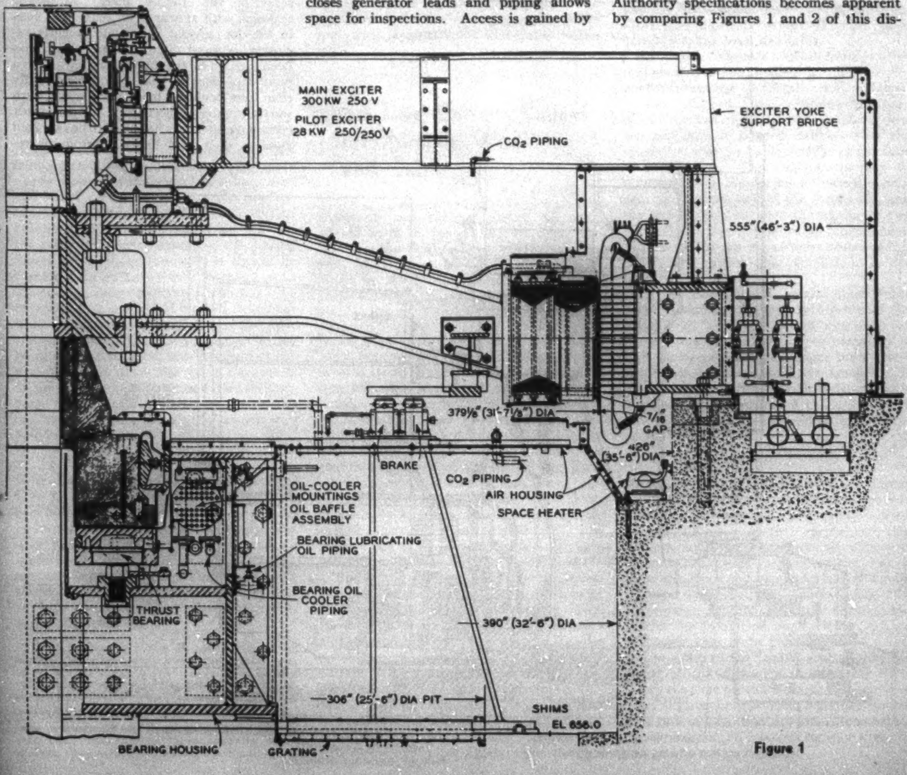


Figure 1

cussion. The latter is a cross section of the 35,500-kva 90-per-cent-power-factor 3-phase 60-cycle 105.8-rpm 13,800-volt umbrella-type generators built by Allis-Chalmers Company for Tennessee Valley Authority's Fort Loudon Station. The Fort Loudon design follows that of the Chickamauga generators except for the thrust block being integral with the rotor shaft and the generator rotor being mounted on and keyed to the top surface of the thrust block. This construction reduces the generator shaft length and simplifies rotor erection and dismantling.

On the other hand the large diameter shaft flanges may introduce difficulties in obtaining acceptable shaft forgings.

In conclusion I want to compliment the authors on their very able paper.

S. B. Griscom (Westinghouse Electric Corporation, East Pittsburgh, Pa.): This paper and its predecessor of the same subject[1] are of great value to the industry. In designing a power station, many decisions must be made between alternate plans which individually will satisfy the basic requirements, yet which differ in intangible respects such as reliability and flexibility. In such instances, the engineers' old stand-by—the economic comparison—is unavailing, and the answer must be obtained by judgment based on experience.

The authors have studied over a period of years many individual problems, resolved them to their fundamentals, and formulated standard practices applying specifically to the Tennessee Valley Authority conditions, but of general value on similar projects. The compilation of these practices will be of great assistance to design engineers for a long while to come by affording comparison of conclusions. Particularly noteworthy is the general use of normal manufacturers standards.

The omission of spare transformers for generating stations and operating more than one generator per high voltage circuit breaker are of particular interest. This reflects two important changes from past practice. First, that transformer and circuit breaker reliability have been improved greatly, and secondly, that systems (including interconnections) have become so large that if a unit is dropped occasionally as a result of apparatus failure, it will not affect service. The unit system as applied to generators is quite a departure from the double-bus double-circuit-breaker generator and high voltage busses with spare transformers which were in vogue some time ago.

The use of automatic reclosing on some of the 154-kv lines is of interest. I would like to know if this has made it possible to defer construction of new 154-kv circuits, or whether its principal use has been to improve reliability of the high voltage network.

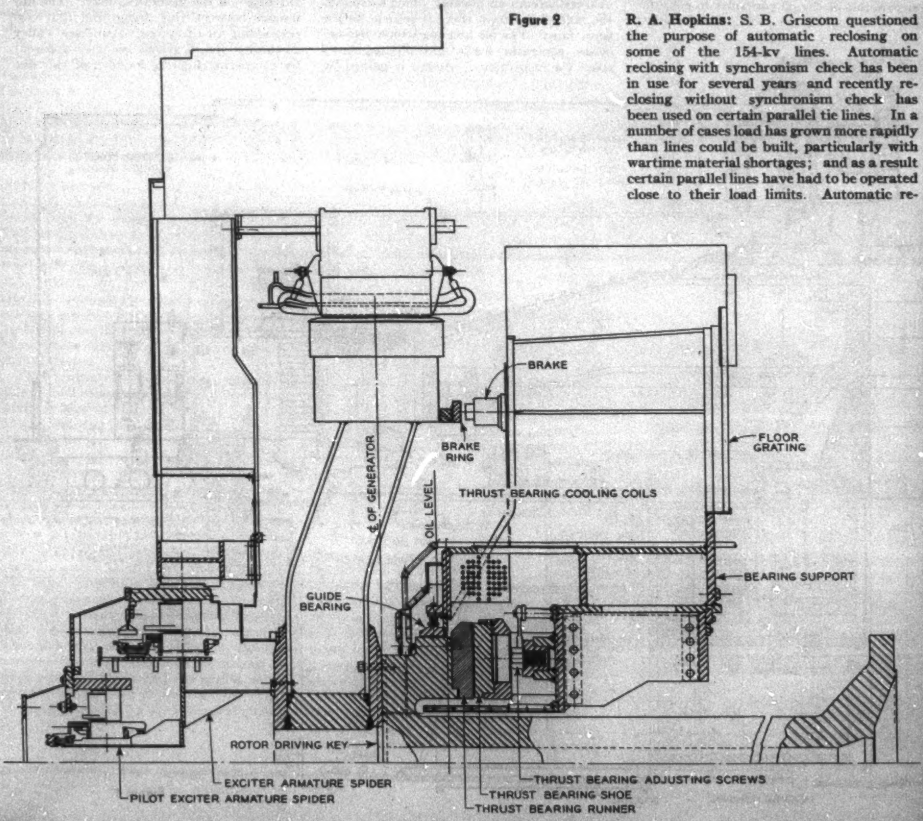
The adoption of an isolating 13.8/12.5-kv transformer-bank for supply of 3,000 kw of village load is an unusual application, wherein the maximum possible protection is offered to the 67,500-kva generators. While increasing the cost somewhat, this procedure no doubt will turn out to have been a good investment during the life of the station.

Reference

1. Tennessee Valley Authority Hydroelectric Stations—Electrical Design, Raymond A. Hopkins. AIEE Transactions, volume 60, 1941, pages 1281–93.

R. A. Hopkins: S. B. Griscom questioned the purpose of automatic reclosing on some of the 154-kv lines. Automatic reclosing with synchronism check has been in use for several years and recently reclosing without synchronism check has been used on certain parallel tie lines. In a number of cases load has grown more rapidly than lines could be built, particularly with wartime material shortages; and as a result certain parallel lines have had to be operated close to their load limits. Automatic re-

Figure 2



closing under these conditions often results in immediately re-establishing parallel operation after a trip-out of one of the lines.

Philosophy of Relaying

Discussion and authors' closure of paper 46-172 by E. W. Knapp and Ben C. Hicks, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 735–40.

Eric T. B. Gross (Illinois Institute of Technology, Chicago, Ill.): I think that all relay engineers should read this paper very carefully and give its contents plenty of thought. The authors have presented the real basis of simplicity in modern relaying, and their conclusions are not restricted in their general application.

It is noticeable that ground fault protection of generators through high impedance grounding and residual voltage relays is recommended, and that the practical experiences have been very satisfactory. There is much reason to give this kind, or some other sort of high resistance grounding, preference in comparison with low reactance grounding of generators. If we would ask ourselves more often, "Why do we ground the neutral of systems and which of the possible answers is to be given preference in each particular case?" many a mistake could be avoided.

It is very interesting indeed, that gas detector relays (indicating the accumulation of gas resulting from incipient faults, and tripping resulting from pressure when more severe internal faults occur) have been used and found very satisfactory. The application of this kind of transformer protection has been wide and successful abroad, but the number of installations in the United States seems to be limited. It should be helpful if the authors would comment more in detail on their experiences with this type of protection, and would include the means which eliminated some of the earlier difficulties mentioned in the paper. It would be of interest to know how many such gas detector relays are in the use in the system of the Shawinigan Water and Power Company, and if other utilities and companies in Canada also use this kind of transformer protection.

E. W. Knapp and **Ben C. Hicks:** In reply to Eric T. B. Gross, the Shawinigan Water and Power Company has in use 40 gas detector relays on transformers and regulators aggregating 600,000-kva capacity, and we understand that other large utilities in Canada use these relays, some more and some less extensively. The chief difficulties that we have encountered in the use of these relays are the following:

1. When an accumulation of gas is released, oil hammer will occasionally cause the pressure element to operate. It has, therefore, been made a rule to block the pressure element whenever releasing gas from the gas accumulation element.

2. Two instances of ruptured diaphragms have been encountered, one in service and one after a shop test. This suggests the advisability of inspecting the diaphragm whenever there is reason to suspect that the relay may have been subjected to excessive pressure.

3. In a few instances, grounds or short circuits have occurred on the relay terminals as a result of moisture accumulating in the terminal chamber. On existing relays this hazard has been overcome by providing a small drain hole in the bottom of the chamber. On new relays the drain hole is provided by the manufacturer and, in addition, the design of the terminal chamber has been improved.

4. Newly filtered oil in a transformer will release trapped air for a period of a few weeks. During this period the relay is used for alarm purposes only.

Our operating experience with gas detector relays extends over the past nine years. Actual operations of the relays during this period have been relatively few. Internal faults on two transformers and one regulator have been detected correctly. In addition to these, a number of miscellaneous conditions, such as low oil, freezing of breathers, and defects in the cooling system have been indicated. There has been no instance of a failure to detect a fault.

Regarding the question of protection in general, our attention has been drawn to the fact that the paper contains no reference to single-phase-switching or Petersen coils. We have had no operating experience with either, but both, when properly applied, undoubtedly have a definite field of usefulness, in clearing ground faults resulting from flashovers on transmission lines without loss of load or loss of synchronism. For this purpose the former is used in conjunction with instantaneous automatic reclosing.

Short-Time Current Ratings for Aircraft Wire and Cable

Discussion and authors' closure of paper 46-145 by B. W. Jones and J. A. Scott, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 644–8.

R. H. Kaufmann (General Electric Company, Schenectady, N. Y.): The results of investigations presented in this paper fill a long felt want in aircraft application technique. The entire practical application significance has been condensed into one simple graph which defines the permissible current-time boundary for the normally encountered variety of conductor sizes over the entire range of current and time. Appropriate test procedures also have been outlined by which other investigators may extend the scope of knowledge. The authors are to be credited with a valuable contribution to the aircraft electrical art.

The results of this work clearly emphasize the need of automatic fault protectors in aircraft electric circuits and make evident the hazards of non-trip-free circuit breakers and manual over-ride features. It can be noted that the flow of a 3,000-ampere fault current will elevate in 0.2 second the temperature of a number 8 copper conductor to the maximum safe value. If current flow be continued for another 0.2 second, violent evolution of smoke will occur, and in less than one second total elapsed time, the conductor will be elevated to the ignition temperature of volatile materials in its insulation. The extremely short time which is available in which corrective action is to taken is entirely beyond the ability of a human being to respond.

A vivid demonstration of the evolution of large quantities of smoke from aircraft cable was performed inadvertently during laboratory tests in which a length of number 4 cable was used to simulate a resistance fault. After several successive intervals of approximately 5,000-ampere fault current flow, each of about one-half second duration, there suddenly occurred a tremendous evolution of smoke throughout the entire length of the conductor which filled the laboratory room and caused several other laboratory attendants promptly to vacate the premises. Once this effect is witnessed, there no longer will be any doubt that such an occurrence positively must be prohibited on aircraft electrical systems in flight.

L. G. Levoy (General Electric Company, Schenectady, N. Y.): This paper should prove to be of substantial value and interest to designers of aircraft electrical systems, and fills a long felt need in the industry. The authors are to be congratulated on providing useful basic data, and pointing the way for obtaining a more rational basis for short-time current ratings of wires than heretofore has been available.

On some aircraft installations, after periods of use, it was observed that the conductor insulation on certain wires appeared to have slipped back, exposing the bare conductor for one-half inch or more near the end, and it seemed unlikely that the insulation was cut back that far at the time of installation. This phenomena may be a result of the longitudinal thermal expansion and contraction of the conductor, and the lack of bond between the conductor and insulation. I would be interested in knowing whether or not the authors observed this effect, and under what conditions it can be expected to appear.

In the zone labeled "smoke starts," the lower boundary apparently represents the lower limit at which smoke was observed on any sample, and the upper boundary of this zone represents the upper limit above which all sample conductors smoked. Is this difference a result of the different types of insulation tested, or of the difficulty of detecting accurately the beginning of visible smoke?

B. W. Jones and **J. A. Scott:** L. G. Levoy mentions the apparent shrinkage of the wire insulation, which takes place most noticeably with heavy current surges and consequent rapid conductor heating. The wire expands rapidly as its temperature rises, it separates from and protrudes out of the insulation which responds more slowly. On subsequent cooling the wire shrinks but carries the insulation back with it, with a net result that looks like slippage.

The width of the "smoke starts" zone is a result of the difficulty of detecting accurately the time at which smoke first starts, or more accurately expressed, reaches a noticeable rate of evolution, for the rate at which smoke is given off changes rapidly with the temperature once it starts.

F-3 Lead Alloy—An Improved Cable Sheathing

Discussion and authors' closure of paper 46-115 by L. F. Hickernell and C. J. Snyder, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 563–9.

Herman Halperin (Commonwealth Edison Company, Chicago, Ill.): For the past 16 years, I actively have been looking for and assisting in the development of a lead alloy for improved cable sheathing. The development of the F-3 lead alloy is, I believe, the first lead alloy that has substantially better quality in all properties than ordinary commercial leads, without having any particular disadvantages. The authors and their associates deserve great praise for the development.

Considerable work on field conditions has been done in Chicago during the past 20 years to avoid premature sheath cracking in manholes incidental to bending caused by variations in daily loading. This work included

1. Enlarging manholes to provide more room for cable movement.
2. Improving the cable training.
3. Modifying the cable fireproofing in the manholes.

The troubles caused by sheath cracks were reduced, but are still a major item. As brought out in an earlier paper,[1] the life of the sheath to cracking in the manhole is in many cases substantially less than the life of the insulation.

Our interest in the development of a new alloy has been accentuated in recent years mainly because we expect to continue the trend of increasing the loading of cables in order to obtain the maximum feasible usage of investment. Recently, the prices of cable have increased roughly one-third. The costs of conduits and manholes have increased even much more in the past seven years. Altogether, such sharp increases in costs emphasize the necessity for carrying heavy loads in underground cable systems, especially since there seems to be no particular tendency to allow electricity rates to increase.

The bending amplitude of the sheaths in the manholes is increased with increased loading. Furthermore, the ability of the sheaths to withstand a given amount of bending is decreased by the higher temperatures at which the sheaths will operate.

We changed a few years ago from belted type cable to shielded type 3-conductor cable for 12-kv service. Then we found that the life of the sheath of shielded cable before cracking in dummy manhole tests was roughly 40 per cent less than for the sheath of the belted type cable. Therefore, we were going forward as far as the insulation is concerned and backwards as far as the sheath was concerned.

We have investigated several types of alloys on an experimental basis, and we are continuing these co-operative investigations with the manufacturers. Two of the alloys that seemed attractive initially subsequently were found to be unsuitable. These two materials age hardened and then had no advantage, and perhaps some disadvantages as far as resistance to manhole fatigue was concerned.

Over the past four years, we have made extensive studies on the F-3 alloy sheath in Chicago and at the University of Illinois. According to our tests, this alloy does not harden with age, and its ductility is retained with age even better than that of chemical lead. Also, the F-3 alloy seems to be suitable for use at considerably higher hoop stresses than ordinary leads in use.

A comparison of the life to fracture of the F-3 alloy to that of ordinary lead in the dummy manhole is shown in Figure 1 for a variety of conditions. All of the tests represented were made on 500,000-circular-mil 3-conductor 12-kv cable of several types having practically the same diameters. In some cases the samples are from the same factory, so that they differed only in type of sheath or binder.

These data indicate that with equal conditions of temperature and movement, F-3 sheaths can withstand about three times the number of bending cycles that copper bearing lead sheaths can withstand. The graph also shows how sharply the life of sheaths decreases when the sheath temperature is increased from room temperature to merely 43 degrees centigrade (110 degrees Fahrenheit). That temperature represents a high average temperature for sheaths in manholes during maximum day-in-and-day-out loading and is the temperature we selected about 12 years ago for some of the research at the University of Illinois.

For some cables an increase of 200 per cent or more in sheath life, as indicated for the F-3 sheath, is quite desirable. For example, the 200-per-cent improvement is essential for some uses of 3-conductor shielded type cable with metallic binder. Based on these laboratory data and related field experience, for equivalent life to sheath fracture for cables having over-all diameters greater than about 1.75 inches, it is estimated that the new alloy will permit an increase of about 25 per cent in daily loading. Some other factors, such as the deterioration of the insulation or system conditions, might reduce the extent to which this advantage can be applied.

These bend tests were made at a rate of about one cycle per minute, whereas the bending in service is at about one cycle per day. In the research on cable sheathing at the University of Illinois, some information is being obtained on the effect of the speed on the life to fracture, for strip specimens are held horizontally at points eight inches apart while the middle is moved up and down. The average number of cycles to fracture with 10-minute cycles for lead was 51 per cent less than the number of cycles with 1-minute cycles, while the decrease was only 37 per cent for the F-3 alloy. Tests with longer cycles are in progress, but results are not available now. If this trend persists to 24-hour cycles in service, then the superiority of F-3 alloy over lead in this respect is even greater than that indicated by the graph.

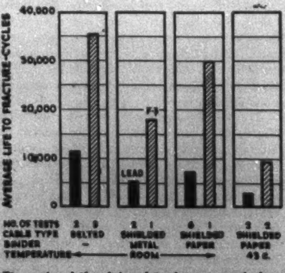


Figure 1. Life of sheath in dummy manhole—0.5-inch movement

Incidentally, our co-operation with the Anaconda Wire and Cable Company in testing and otherwise investigating the F-3 alloy started six years ago. We did not place any commercial order for cable with the F-3 alloy until November 1944. These time periods are mentioned to indicate that the investigations of the suitability of a new lead alloy take years.

REFERENCE

1. LOAD RATINGS OF CABLE, Herman Halperin. AIEE TRANSACTIONS, volume 58, 1939, October section, pages 535–52.

C. T. Hatcher (Consolidated Edison Company of New York, Inc., New York, N. Y.): The authors have presented data on a subject which is of extreme importance to the industry at the present time. A lead sheath is being offered which apparently answers many of the disturbing problems which we are encountering today. If field experience confirms the laboratory test data, then a great contribution has been made.

The authors list oil filled cable operating at 10 to 15 pounds pressure as one of the places where this type of lead could be used advantageously. A more specific requirement for the use of this lead in oil filled cable systems is where the pressures may be in excess of 15 pounds, and therefore double sheath cable is required. Elimination of double sheath cable is one of the things which we always attempt to do in the initial design of an oil filled system. I have attempted to obtain this end during the past number of years, and if the new lead will answer this problem a considerable contribution has been made.

A lead sheath which will give longer life under loading conditions causing expansion and contraction of cables is needed. We have had several experiences in recent years where the cable sheath has been broken on gas and oil filled cables because of cable movement in manholes where proper racking conditions could not be obtained, and if a sheath is available that will help this situation it certainly will be helpful.

Abe Tilles (Pacific Gas and Electric Company, San Francisco, Calif.): As the authors state, improvement in dielectric of cables has made the lead sheath more and more of a limiting factor in the life of the cable. Accordingly, the very fine qualities reported in the paper for the lead alloy described are of great interest to all users of lead covered cable.

It would be of interest to know how much of such sheathing has been placed in commercial use and which operating companies have made major installations. It also would be of interest to know if this sheathing has been, or is being, applied to higher voltage, for example, 120-kv cable.

T. H. Haines (Boston Edison Company, Boston, Mass.): During the past few months, the Boston Edison Company has been testing F-3 lead sheath at its own laboratory. Three 10-foot samples of 3-conductor 350,000-circular-mil cable having 260 mils of paper insulation and 125 mils of F-3 lead were bent to a radius of 42 inches. One end was fixed and the other was moved back and forth by a motor driven arm which made ten 2-inch strokes per minute. An air valve and pressure relay were attached to each test specimen. During the test the cable sample operated with an internal air pressure of five pounds per square inch. When the sheath cracked, the relay operated, stopping the test. The number of cycles of bending was read on a counter.

For comparison, three samples of cable sheathed with commercial or copper bearing lead were tested in the same way. These were chosen to have as nearly as possible the same weight, outside diameter, and lead thickness as the Anaconda Wire and Cable Company sample. One was a piece of 3-conductor number-4/0 15-kv split sector cable and the other two were cut from a length of 3-conductor 350,000-circular-mil 28-kv type-H cable. The number of cycles required for failure with the F-3 samples were 20,260; 25,695; and 19,484. The number of cycles required for failure with the control samples were 9,931; 6,061; and 8,307.

In each case it was noted that the F-3 lead cracked between the bulges developed by the test, while the other sheaths cracked on the bulged surface.

Another type of test was tried which consisted of placing strips of sheath between strips of brass and clamping the assembly jaws six inches apart. The centers of the strips were vibrated with a 7/32-inch stroke at a frequency of 240 cycles per minute. We have not had this device in use long enough to have established a standard reproducible test, but in the few runs which were made the strips of F-3 lead remained intact longer than the other samples.

Howard S. Phelps (Philadelphia Electric Company, Philadelphia, Pa.): Physical tests have been made on limited samples of F-3 lead alloy cable sheathing and sheathing pipe by the Philadelphia Electric Company.

INTERNAL PRESSURE HYDRAULIC BURSTING TESTS

Time-to-Burst. Hydraulic bursting tests at constant internal pressure[1] at 110 degrees Fahrenheit on sample of F-3 lead alloy sheathing pipe indicate this alloy has characteristics quite different from those of lead sheathing alloys heretofore tested. The time-to-burst versus initial stress data substantially corroborate the F-3 curve of Figure 5 of the authors' paper for stresses of 2,400 and 1,600 pounds per square inch. Present indications are that the time-to-burst for a specimen of approximately two inches internal diameter and 9/64-inch nominal thickness may be of the order of 40 years at 100 pounds per square inch internal pressure (720 pounds per square inch initial stress). This contrasts with a number of other alloys tested, none of which appears to be very much longer lived than commercially pure lead at stresses of the order of 300 to 400 pounds per square inch.

Type of Failure. It has been found characteristic of lead sheathing having appreciable percentages of alloying elements (that is, more than one-tenth per cent), that the bursts occur in the welds. The F-3 sheathing pipe appears to be no exception. The failures of the specimens tested at 2,400 and 1,600 pounds per square inch initial stress occurred in die welds.

Total Stretch. The total stretch of F-3 sheathing at failure, as indicated by the authors in Figure 9, increases with decreasing stress, but appears to be of the order of 15 per cent increase in diameter. In the Philadelphia Electric Company tests, the total stretch for the specimen stressed at 2,400 pounds per square inch was 2.8 per cent, and 4.0 per cent for the specimen stressed at 1,600 pounds per square inch. This corroborates the tendency toward increasing total stretch at failure with decreasing stress, indicated by the authors. However, in the Philadelphia Electric Company tests the total stretch was considerably less than for copper lead or chemical lead for the same stresses, indicating the F-3 sheathing is less ductile at high stresses.

In 66-kv single conductor solid type cables, the pressures developed by thermal expansion can become very great. Although A.S.T.M. grade leads (not alloyed) are quite ductile at high stresses, a considerable proportion of failures of such cables are attributed to sheath bursts by internal pressure, despite the relief afforded by stretching of the sheath. It would appear that for this type cable, the reduced ductility of F-3 lead alloy at the higher stresses may result in more internal pressure failures.

BENDING RESISTANCE

Philadelphia Electric Company has made limited dummy manhole bending tests[2] on cables sheathed with F-3 lead alloy. The results indicate that the life of cable sheathed with this alloy is of the order of three times that for the same type of cable sheathed with the types of lead generally used commercially.

HARDNESS

Hardness tests of F-3 alloy lead indicate it to be intermediate in hardness between Asarco (0.26 per cent tin, 0.03 per cent magnesium, 0.025 per cent calcium, and 0.02 per cent bismuth) and the leads in general commercial use.

REFERENCES

1. INTERNAL HYDRAULIC BURSTING TESTS OF LEAD CABLE SHEATHING, H. S. Phelps, A. M. Gates, F. Kahn. *Proceedings*, American Society for Testing Materials, volume 40, 1940, page 885.

2. CYCLIC MOVEMENT OF CABLE—ITS CAUSES AND EFFECTS ON CABLE SHEATH LIFE, Clement S. Schifreen. AIEE TRANSACTIONS, volume 63, 1944, pages 1121–30.

G. B. McCabe (Detroit Edison Company, Detroit, Mich.): The new F-3 alloy of the Anaconda Wire and Cable Company appears to have many desirable properties that appeal to utility operators of cables, who are always glad to welcome and adopt any proved new development that insures longer and more troublefree life for cable. If actual field experience under a variety of operating conditions supports the test results and the many claims made for the new alloy, there undoubtedly will be a demand by the operators for the more universal use of it as sheathing of cables to be installed on their systems.

The primary purpose of my discussion is to correct an erroneous conclusion which the authors drew from reference 9 of their paper.

In using this reference, the authors stated that calcium alloys (0.025 to 0.031 per cent) are difficult to extrude (very drossy) and undergo slow and serious age hardening, which has caused troubles to develop in a few years of service due to brittleness and early cracking in the manholes.

The authors have used the term calcium alloy, whereas a distinction should be made between straight calcium lead cable sheathing and other alloys of lead containing calcium. I wish to make it clear that we have had no adverse experience with straight calcium lead sheathing in five years of service.

The article referred to dealt principally with trouble experienced with a quinary alloy composed of magnesium, tin, bismuth, calcium, and lead, which was designated as alloy B. Cable sheathed with this alloy, which originally was installed in 1940, developed innumerable cracks through the sheath upon being reinstalled. A large circumferential crack also was discovered in the expansion loop of an aerial line of cable sheathed with this alloy. A similar crack was uncovered in a slight offset two feet from a joint in a heavily loaded underground cable.

Laboratory examinations of sections of the sheath indicated that the failure of the sheath was caused by age hardening, with resultant loss in ductility as manifested by a marked intercrystalline brittleness and decrease in bending strength.

Also in 1940 there was installed a considerable quantity of cable sheathed with calcium lead containing a nominal content of 0.027 per cent of calcium.

This cable, which was of the 350,000-circular-mil shielded 3-conductor 24-kv type, was installed in various parts of the system in order to obtain operating experience under a variety of conditions, such as on aerial construction, on the feed from a large base-load generator to a switching station, and in cable transmission lines subjected to heavy load cycles.

During the service of more than five years for this cable, there has been no sign of distress in the sheathing at any point. Samples taken from a length, on which three installations were simulated, revealed no evidence of even surface checking of the sheath.

Figure 2 of this discussion indicates the influence of calcium content on tensile strength. It also shows the relatively short time necessary for calcium alloys to approach their ultimate tensile strength. The calcium lead (0.025 per cent) starts out with a tensile strength of about 3,000 pounds per square inch (curve 2) and attains values in excess of 3,600 pounds per square inch within 20 days (curve 3). Alloy B, however, required about one year's aging before it approached its ultimate value (curve 1). The

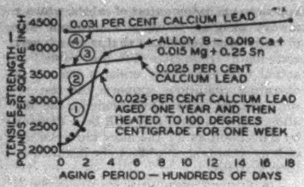


Figure 2. Increase in tensile strength of alloy sheath after aging

Samples represented by curves 3 and 4 were 21 and 48 days old before the first test was made

data for the calcium lead, which incidentally were received largely from the Anaconda Wire and Cable Company laboratory, do not agree with the curve on Figure 1 of the present paper, which shows that it takes almost two years instead of a few weeks for 0.025-per-cent calcium lead to attain a tensile strength of 3,600 pounds per square inch.

Figure 3 of this discussion shows the uniformity of bending strength with time for calcium lead, compared to the large decrease for alloy B in less than one year.

In those tests reported in the paper by Hickernell and Snyder, in which data were obtained on calcium lead composed of 0.025 per cent calcium, there appeared to be but little difference between it and the F-3 alloy, particularly as shown in Figures 2, 4, and 12.

Metallurgical consideration indicates that the calcium content could be reduced to 0.020 per cent and thus improve the important property of ductility without significant reduction in tensile strength, and resistance to creep, fatigue, and abrasion. A 0.020-per-cent calcium lead probably would have a tensile strength of about 3,200 to 3,300 pounds per square inch, approximately that of the new F-3 alloy.

While I am in no position to question the statements of the authors with reference to extrusion difficulties encountered with calcium alloys, their data fail to support the implication that poor service is an inherent quality of calcium lead cable sheath.

Data on copper bearing lead reveals tensile strength of about 1,950 pounds per square inch. It would appear that F-3 alloy, which has a tensile strength of 3,250 pounds per square inch according to Figure 1 of the paper, would be at least 65 per cent stronger than copper bearing lead instead of "not much stronger" as stated in the paper. Also, the Brinell hardness number of F-3 alloy is almost double that of copper lead.

On the occasion of previous developments by a cable manufacturer, such as compact stranding, carbon dioxide cleansing of impregnating tanks, supercalendered paper insulation, shielded construction, gas filled cable, and the like, agreements have been effected between the various manufacturers concerned for the incorporation of such new developments in cables produced by them. What is the probability of the new F-3 alloy becoming available soon to the other cable manufacturers so that the users may benefit from the claimed and proved advantages of such an improved sheathing on their future cable purchases?

How will the cost of cable sheathed with F-3 alloy compare with those sheathed with copper-bearing lead and the alloys of tin or antimony?

REFERENCE

1. AGE HARDENING PROVES FATAL TO LEAD-ALLOY CABLE SHEATH, G. B. McCabe. *Electrical World*, October 13, 1945.

C. R. Bond and **F. W. Nimmer** (Ohio Edison Company, Akron, Ohio): Hickernell and Snyder are to be complimented on the excellence of their paper. There appears to be a substantial contribution to the reliability of electric power cables by the development of a new lead alloy to be used in cable sheathing. They have tried to take into consideration all of the essential requirements for lead sheath demanded by the cable users at this time. It also is interesting to note that a good deal of time seems to have been devoted to the accompanying research problems in order that essential properties might be given adequate consideration, prior to the presentation of a definite recommendation of superiority of any given type of cable sheathing. Not only has the technical research work been done, but actual commercial installations have been made which have been in service for a period of several years in particularly critical locations, in order to determine the variations between laboratory test and practical commercial installations.

In the interest of completeness, we suggest that two additional items be considered; namely, any new alloys used for lead sheathing should not affect the insulating properties of the various insulating materials to be employed, and, also, such lead sheathing should have good thermal conductivity. Up to the present time, we have not been concerned particularly about these features, but as new alloy sheathings are developed, we must be certain that deterioration of the cable will not result from a lack of recognition of these features.

If this new F-3 lead alloy fulfills the requirements as well as we hope that it will, Hickernell and Snyder will have made a very substantial contribution to the cable art.

Earle E. Schumacher (Bell Telephone Laboratories, Murray Hill, N. J.): I have read with interest the paper by Hickernell and Snyder and consider the work which they have done a valuable contribution to the cable sheath art. My major comment concerns the authors' statements regarding lead alloys containing 0.025 to 0.031 per cent calcium. The embrittlement behavior cited in their reference 9 actually concerns a complex lead calcium alloy containing magnesium rather than the binary lead calcium alloy. It is well-known that magnesium is likely to induce brittleness in lead. The paper by G. B. McCabe emphasizes that the binary lead calcium alloy is not subject to this embrittlement and states that ". . . a reel of lead calcium alloy sheathed cable, that had been installed and operated for $2^1/_4$ years and later removed from the duct, was subjected to re-reeling twice at the cable plant. No evidence whatsoever of even surface checking of the sheath was found."

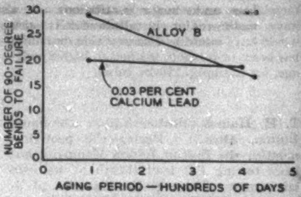


Figure 3. Decrease in bend strength of alloy sheath after aging

These sheaths were extruded in the latter part of April 1940, but were not tested until nearly three months later

This confirms Bell system experience with lead alloys containing 0.025 to 0.031 per cent calcium. Hundreds of miles of telephone cable sheathed with this alloy have shown no evidence of failure because of embrittlement.

With regard to the authors' statement about dross in lead calcium sheath, microscopic examination of a great number of samples of the binary alloy have shown this type of sheath to be as free from dross as other types of lead alloy sheath extruded on intermittent presses.

My other comments deal with the interpretation of creep and fatigue data in general. Creep tests at the Bell Telephone Laboratories carried out over a period of eight years have shown that data taken at stresses of 400 or 500 pounds per square inch are not necessarily a reliable indication of the behavior of the same alloys stressed at 100 or 150 pounds per square inch. The slope of the stress-rupture curve often changes at the low stresses, and this may change the relative rating of the alloys.

With respect to fatigue data on lead cable sheath alloys, we have found as a result of extensive studies that frequently the relative merit of different materials is dependent upon the speed of testing. Since the major deflections of sheath in service are in most instances of the order of one or two cycles per day, there is a substantial reason for testing at speeds much lower than 700 cycles per minute. Accordingly, most of our tests now are made at a speed of 15 cycles of reversed deflections per hour. Also, since the forces which produce cyclic sheath deformations in service are so great that sheaths regardless of their properties tend to be deflected similar amounts, there is considerable merit in making fatigue comparisons on a deflection cycle basis rather than on a stress cycle basis.

G. B. Shanklin (General Electric Company, Schenectady, N. Y.): F-3 alloy is undoubtedly a step forward in the art of cable sheath construction. It is one of a group of arsenical alloys, all of which have similar characteristics. There are a number of relatively hard high strength lead alloys available for cable sheathing. Most of these are age hardening but some are age softening (self-annealing). The arsenical alloys belong in the latter class, and this explains their higher bending fatigue life, which is their outstanding characteristic.

Any type of hard high-strength lead-sheath alloy requires exceptionally close metallurgical and foundry practice control to assure that neither age hardening nor age softening will become objectionable or dangerous after a long period of service. In underground power cable work, the two most important characteristics that need control throughout the life of the cable are bending fatigue and long-time creep strength. These two characteristics oppose each other, however. High bending fatigue resistance requires an age softening alloy, while high creep strength resistance requires an age hardening alloy. The obvious goal is to obtain an alloy that neither age softens nor age hardens but retains complete stability of characteristics throughout its life. Everyone who has had experience with the high strength alloys knows how difficult it is going to be to reach this goal. With this in mind, it might be well to review briefly the major sheath requirements for the different types of high voltage cable used in this country.

For solid type cable, especially up to 18-kv rating, resistance against sheath fatigue caused by cable movement is an outstanding requirement with high creep strength a lesser requirement. Practically all sheath fatigue troubles have occurred in this type and voltage range of cable and are primarily caused by high operating temperature range and small manholes with inadequate bends and offsets. As the voltage rating of solid type cable increases, bending fatigue becomes of lesser importance and creep strength of greater importance, because of the fact that the operating temperature range decreases, and more liberal bends and offsets in the manholes generally are used. Also, oil filled joints and reservoirs increase the importance of high creep strength.

During the past 19 years, something like 4,000,000 feet of standard low pressure oil filled cable have been installed in this country. As far as sheath failures caused by cable movement are concerned, the record is remarkable, in fact practically perfect. Most of this cable has copper bearing single sheath, and this service record plainly indicates that there is still plenty of latitude for load and cable movement increase before there will be any serious problem of bending fatigue. The reason for this is that the manholes always have been large to accommodate the accessories, and the bending radius and offsets always have been quite liberal. This experience shows that there is no real need of a high bending life alloy as far as oil filled cable is concerned. There is some need, however, for a high creep strength alloy sheath for oil filled cable, particularly on steep grades where hydrostatic head pressure builds up. The economic over-all saving that this would represent, however, is much less than most engineers realize. For average profiles usually encountered, it would mean a few less stop joint manholes. This in turn would in average cases mean only a few per cent saving in over-all cost. In return it would mean a definite commitment to high hoop stress operation of the sheath. Hickernell and Snyder are not exactly correct when they say that copper bearing sheath is limited to a hoop stress of 125 pounds per square inch. This limit actually applies to two-per-cent tin alloy sheath at full operating temperature. Long experience has shown that copper bearing sheath can be worked safely to a hoop stress of 150 pounds per square inch without noticeable creep in service. To show any appreciable economic gain, an alloy sheath would have to be used which would allow a hoop stress of at least 300 pounds per square inch, and resistance to creep under this stress must be retained throughout the life of the cable. In the case of age softening alloys such as F-3, we would have to be absolutely sure that these limits could be met. There would be little doubt about it with age hardening alloys such as Asarco, but it is yet to be proved with age softening alloys, where the relative decrease in creep strength with increase in time and temperature is more marked.

With low pressure gas filled cable, there has been a limited amount of trouble with sheath fatigue caused by cable movement, but this has been only at 15 kv and for the reasons already outlined for similar trouble with solid type cable. This type of cable is used under much the same conditions as solid type, and there is no question but that high bending fatigue life is desirable at the lower voltage ratings. High creep strength is even more important, however, since the present low pressure gas filled cable with copper bearing sheath is limited to a working gas pressure in the order of 15 pounds per square inch. If this could be doubled, it would allow a reduction in insulation thickness and actually become a medium pressure system. It would, however, mean that a hoop stress in the order of 300 pounds per square inch would have to be accepted. There is an open question whether an age softening alloy could withstand this working stress indefinitely at full allowable operating temperature and with the cable movement that will be superimposed. So far, we have used only the double reinforced type of sheath for medium pressure systems. We have hesitated to offer Asarco sheath unconditionally for this purpose, not because there was any doubt about its permanently high creep strength, but because of its tendency to age harden. For the present at least, we would hesitate to offer an age softening alloy for reverse reasons, and feel that a good deal more must be learned about the stability of long-time creep strength before such extreme limits of hoop stress are adopted on a wide scale. Until these points are cleared, the logical field of use for F-3 alloy is in taking advantage of its bending fatigue resistance as required by solid type cable.

A systematic research program is being carried out by our laboratories, and presumably by others. Based on results to date, we feel that it is only a question of time before improved lead sheath alloys having high creep strength stability become available.

R. J. Wiseman (The Okonite Company, Passaic, N. J.): Hickernell and Snyder have presented a most interesting paper on a lead alloy which they have developed to meet the problems which confront us today in the operation of impregnated paper cables. They list the factors which determine if a lead is good or bad. All of us for several years have been studying lead alloys as we have come to recognize that to meet the demands of the electrical industry for a long life lead sheath, it is necessary to turn to alloys. Some have been promising in the beginning, but because of age hardening have failed to prove in. Others are good in some physical respects, yet fail in other respects, so we have been hesitant to try them out on a large scale. It appears that the authors have a very promising alloy and even though their tests give every indication of successfully meeting the all-around physical properties of good sheathing, we all know that field experience will be the final criterion after a large production of pipe with the variability of foundry practice present.

There is one interesting fact to me which my company, in its study of this subject, found to be significant. You will note that the authors do not use a large amount of an alloying metal, but small amounts of several metals. We found this to be favorable in our studies. We have studied the same metals as the authors report and also others, but in all cases small amounts were used. In some cases there seems to be a limit to the amount which is helpful, beyond this the alloyed lead loses its promising properties. There are some metals, even in small amounts, which are undesirable. As many of us have been working on this problem for years, we can expect a disclosure, from time to time, of other alloys of lead also approaching the ideal sheath to meet the increasing demands of the utilities for higher operating temperatures without sheath failures, which is the major cause of cable failures today.

L. F. Hickernell and C. J. Snyder: In general, the experimental data presented by the discussers corroborate data in the paper. In evaluating the several properties and in the possible applications of the new alloy, some differences in opinion exist. This is to be expected with any new material and serves to indicate the value of free and open discussions on engineering matters as afforded by the forums of professional societies.

Specifically, in reply to the discussers, we offer the following:

(*Halperin*) We greatly appreciate the encouragement of Halperin in this development and are indebted to him and his engineering associates, the Commonwealth Edison Company, Utilities Research Commission, and University of Illinois for their painstaking research program and evaluation of this new alloy. We hope at some future time he will present more of these interesting and valuable data.

Although not strictly on the subject of the paper, we cannot refrain from pointing out in his figure the increase in life on bending obtained on shielded cable by replacing the metal binder with a paper binder. This superior mechanical performance is obtained without any sacrifice in electrical characteristics by constructing the binder of carbon black tapes which have transverse (radial in the cable) as well as longitudinal conductance and hence form a short effective conducting path from the individual shielded conductors to the sheath.

(*Haines*) On the basis of average values, we note that the life of F-3 alloy was approximately three times that of the copper lead sheath, thus verifying results obtained by others. This confirmation is of special interest because the bending of the cable samples was conducted on a different type of test equipment.

(*Hatcher*) As pointed out in the paper, the higher working stress of F-3 alloy will permit the use of single instead of double sheath in certain installations. We hope that a trial installation can be made to study its operating performance in this service.

(*Tilles*) Up to August 1, 1946, about 140,000 feet of cable (or 600,000 pounds of alloy) are on order or have been shipped to Commonwealth Edison Company, Kansas City Power and Light Company, Duquesne Light Company, Consolidated Gas Electric Light and Power Company of Baltimore, Pittsburgh Plate Glass Company, Detroit Edison Company, and Arkansas Power and Light Company. The highest voltage cable rating to date is 45 kv.

(*Phelps*) With reference to bursting strength, we have experienced no failures in over four years of F-3 alloy at stresses 1,000 pounds per square inch or lower in samples on test. Other alloys tested are not appreciably better than copper lead at low stresses (see Figure 5). We are glad to note confirmation of this desirable performance.

In our experience, bursts occur at the thinnest portion of the specimen, which is quite likely to be in the region of the top or bottom weld. This is a limitation of present extrusion equipment rather than an inherent property of any material. Despite this, reference to Figure 5 indicates that at 1,600 pounds per square inch time-to-burst of F-3 alloy is approximately 500 times that of copper lead.

In discussing inherent properties of these materials, it is important to note that most alloys are characterized by a difference in the grain structure at the weld areas compared to the remaining structure, which is the cause of their relatively low bursting strength. However, F-3 alloy has uniform grain structure throughout (see Figure 4 of this discussion). Accordingly, F-3 alloy inherently has the same unit strength in the weld areas as elsewhere in the structure.

As shown in Figure 9, F-3 alloy is less ductile than copper lead at the same stresses. However, in the case of 66-kv solid type cable, operated with oil reservoirs, we do not see quite how the use of a material with greatly superior bursting strength would lead to more internal pressure failures.

(*McCabe*) In view of his favorable operating experience with 0.025 per cent calcium lead alloy, it would appear that our reference 9 is in error. However, it is well-recognized that calcium lead does age harden as indicated by McCabe's curve 2, Figure 2 of his discussion. This age hardening ultimately lowers the bursting strength and bending resistance.

Curve 3 (obtained from our laboratory in 1941) in McCabe's Figure 2 was based on data from cable sheath. Our Figure 1 is based on tubes. The mechanical working in tightly applying the sheath onto the cable core rapidly ages an age sensitive material. Consequently, the initial tensile values on a sheath are approximately the same as those on a tube after a considerable period of aging.

Referring to McCabe's Figure 3 of his discussion, the principal purpose of the IPCEA* bending machine is to check quality at the factory by disclosing poor welds, lamination, and so forth. We believe it is not sufficiently sensitive to differentiate aging characteristics between materials.

As stated, 0.025 per cent calcium lead is about the same as F-3 alloy in our Figure 2 (creep resistance), Figure 4 (creep rate), and Figure 12 (vibration fatigue). However, after aging calcium lead lacks bending resistance and bursting strength (Figure 5) of importance in power cable service.

The licensing of other manufacturers to use this patented cable sheathing material is a matter of negotiation between commercial departments.

Based on present market conditions, it is expected that F-3 alloy will cost no more, and perhaps somewhat less than two per cent tin or three-fourths-per-cent antimony alloy. Obviously, the exact differential between any of these alloys is dependent upon the metal market at the time of quotation.

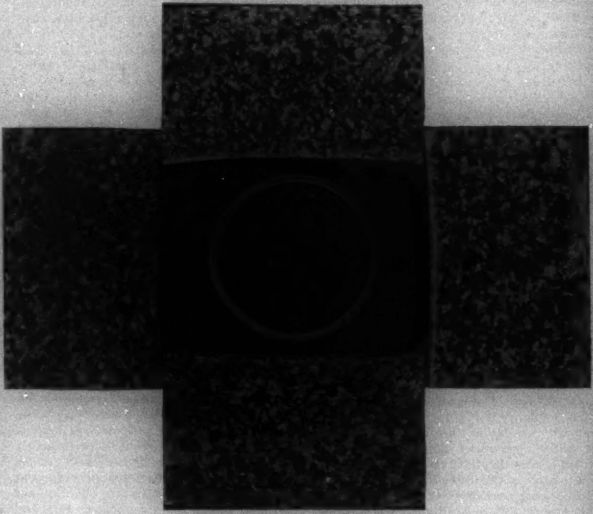


Figure 4. Structure of F-3 alloy

Upper and lower microphotographs show weld areas

Two side microphotographs show nonweld areas

(*Bond and Nimmer*) With reference to thermal conductivity, we have made no determinations, but we would not expect it to differ appreciably from other lead sheathing material. In current-carrying-capacity calculations, thermal resistance of the lead sheath is assumed to be negligible. Thermal emissivity depends primarily upon the condition and color of the sheath surface and has been standardized as one value for all sheathing materials.

Regarding the effect on the insulation, we find that the chemical composition of the sheath material makes little difference in the formation of lead soaps. Of course, in our type CB cable design, any formation of copper or lead soaps is neutralized completely by the adsorption properties of carbon black.

(*Schumacher*) Our statement regarding the effects of age hardening of calcium lead alloy was based primarily upon the results of dummy manhole tests and some operating experience in power cable service. Apparently, our reference 9 to the McCabe paper was unfortunate. We never have questioned the adequacy of 0.025 per cent binary calcium alloy for telephone service. As a matter of fact, until the development of our new alloy, we recognized this calcium alloy as the equal or superior to the customary alloys used for power cable service.

When melted, calcium alloy forms surface dross much more readily than other alloys because of oxidation of the calcium. This is undesirable from an extrusion standpoint because

1. The dross, if not thoroughly removed, may be extruded inadvertently in the finished sheath.
2. The calcium content must be replenished frequently.

Consequently, additional process and metallurgical control is required. This can be accomplished, as demonstrated by Western Electric Company and our plants, but at increased cost. F-3 alloy does not dross readily, in fact it forms less dross than copper lead, and its chemical composition remains stable when melted.

In our paper we have designated high frequency test results as vibration fatigue and low frequency as bending resistance. We agree that the latter is the more important for power cable in ducts. In his discussion of our paper, Halperin has reported his results on 1-minute (60 per hour) and 10-minute (6 per hour) cycles in addition to his dummy manhole tests at 70-second (51.4 per hour) cycles.

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* Insulated Power Cable Engineers Association.

(*Shanklin*) In assuming that F-3 alloy is age softening because it contains arsenic, Shanklin apparently has overlooked the fact that it also contains tin and bismuth. In over four years of testing, we have found no age softening except for a very slight amount in the first few months (see Figure 1). F-3 alloy is stable in properties, purposely so designed, and, in our opinion, meets the goal of "an alloy that neither age softens nor age hardens but retains complete stability of characteristics throughout its life".

We do not agree that "high bending fatigue resistance requires an age softening alloy, while high creep strength resistance requires an age hardening alloy". Calcium alloy has comparatively good bending resistance and antimony alloy has poor creep resistance, yet both are age hardening materials.

Practically all power cables undergo daily load cycles and hence are subjected to bending. A sheath which has higher bending resistance should improve cable life (or conversely, permit higher loading for the same life) whether solid, oil filled, or gas filled type. Whether or not the additional cost of such an alloy sheath is justified economically depends upon the individual installation and operating duty.

With respect to oil filled cable, may we suggest that present specification operating temperatures have been selected with the limitations of previously available sheath materials in mind which has contributed to the good operating record of present installations.

Experimental data indicate that copper lead sheath can not be operated safely at 150 pounds per square inch at *maximum permissible temperatures* for normal and emergency service. This is recognized in the AEIC† specifications which limit the stress to 125 pounds per square inch for standard sheath thicknesses. For cases where the stress exceeds 125 pounds per square inch, the thickness of sheath and composition of lead should be as agreed mutually between purchaser and manufacturer.

We have reason to believe that very few oil filled cables in this country actually are operating at 150 pounds per square inch; also, that most of them are operating at moderate temperatures. In several installations, when loaded to permissible temperature limits, sheath cracks have developed.

At the present time, we have suggested 200 pounds per square inch as a conservatively safe working stress for F-3 alloy. We anticipate that operating experience will indicate that this value can be increased.

Granted that this stress would not permit a reduction in insulation thickness of low pressure gas filled cable, we believe the use of F-3 alloy sheath would improve its performance by reducing the sheath troubles which have been experienced in the field with this type of cable.

(*Wiseman*) We are pleased to have the chairman emphasize the importance of small amounts of alloying elements on the properties of sheathing materials and to point out that an alloy is required to improve cable performance.

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† Association of Edison Illuminating Companies.

Fundamental Properties of the Vacuum Switch

Discussion of paper 46-126 by R. Koller presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 597–604.

Jerome J. Taylor (The Detroit Edison Company, Detroit, Mich.): Perhaps everyone has speculated to himself about vacuum breakers, and there has been much experimental work on them. Comparatively little has been published. The author develops the problem in an incisive manner, and derives definite results on many interesting points. The paper itself lacks a few statements as to conditions under which the tests were performed. Curves show current points up to 100 amperes, but oscillograms are all in the 10- to 25-ampere range, and corresponding values of test circuit voltage and inductance are not presented. From a design standpoint, factors that cause failure of a switch to interrupt, and sequence of events when it does not, are of first importance. These are not discussed with the same attention as are the successful operations.

The small contact separation required in vacuum breakers raises the question whether for a-c service they could not be timed to open at current zero and be separated fast enough to prevent subsequent breakdown of the gap. There then would be realized the dream of opening high-power circuits without release of explosive energies within the breaker. If internal breakdown strength of a high vacuum switch is assumed to be of the order of 10^6 volts per centimeter separation of the contacts (limited by cold-cathode emission), an opening speed of 10^2 centimeters per second results in a nominal recovery-voltage strength of 10^2 volts per microsecond. This is not necessarily the limit in contact speeds and begins to compare favorably with many actual circuit characteristics.

T. E. Browne, Jr. (Westinghouse Research Laboratories, East Pittsburgh, Pa.): Doctor Koller has presented some experimental results obtained with a vacuum switch in a relatively low power d-c circuit. While these are of considerable scientific interest and are of value for the design of low power switches such as vacuum relay contacts, a word of caution seems to be needed against trying to extrapolate these results to the design of a high power vacuum circuit breaker. First, as is well-known, the problems involved in interrupting large alternating currents are such that studies of the interruption of small direct currents are applicable to them in only an indirect way. In stating that the literature contains no explanation for the lack of success of vacuum switches in power circuits, Doctor Koller apparently has overlooked such an explanation given by Dr. Slepian in a paper originally presented at the International Electrical Electrical Congress in Paris in 1932.[1] In this paper the action of the vacuum switch in interrupting alternating currents is analyzed briefly but completely. The principal difficulty is shown to be the finite thermal lag of the vapor emission from the electrodes, particularly but not exclusively the cathode.[2] In spite of the high rate of dispersion (not true diffusion) of vapor in an evacuated vessel, this "after-boiling" results in a considerable vapor density between the switch contacts at and just after a current zero. Under favorable conditions the restored voltage causes breakdown in this vapor only to a glow which loses its conductivity with the subsequent disappearance of the vapor blast. The weakness of the switch is, however, that such a glow shows a tendency, increasing with vapor density, current, and voltage,[3] to change over suddenly to an arc. Under such conditions the probability of this glow-arc transition (similar to the arc-back of a mercury arc rectifier) is high, and the switch is not a reliable interrupter.

Doctor Koller, however, in studying d-c interruption has focused his attention on the opposite phenomenon, the transition of an arc to a glow. His oscillograms, like those of some other investigators,[4] show many rapidly repeated transitions back and forth between arc and glow with the arc-to-glow transition finally "winning out", so to speak, when his interruptions were successful. This arc-to-glow transition tendency, or arc instability, is actually a source of trouble in an a-c breaker since, if it occurs too long before a normal current zero (at too large a current) in an inductive circuit, it can produce troublesome and possibly dangerous high voltage surges of very steep wave front. In the commonly-used low-current vacuum contacts, capacitor shunts often are required to absorb these surges.

Even if the delayed vapor blast could be controlled or eliminated entirely, the application of the vacuum switch to high voltage circuits still would face the difficulty of the apparently inherent lack of reliability of an evacuated space as an insulator which gives rise to random arc-back even in high vacuum thermionic rectifier tubes operated at high voltage. It appears that the glow-arc transition retains, at high voltage gradients, a finite probability of occurrence even at vacuum pressures where glows are invisible and so ordinarily are not imagined.

I should like to ask Doctor Koller for a fuller explanation of his suggestion of the use of multiple contacts for interruption of higher currents and voltages with vacuum switches. It seems very doubtful to me that contacts can be operated in parallel to increase the current interrupting limit, since, even with perfect contact synchronization, separate arcs ordinarily cannot exist in parallel with each other. A series arrangement of contacts undoubtedly would improve the voltage interrupting ability provided the voltage be distributed evenly among the breaks. Furthermore, experiments with the effect of series operation on the arc-back probability of mercury vapor rectifiers[5] indicate that the reliability of a series connection of breaks might be very much greater than that of a single break at the same per-break voltage. This follows from the observed random nature of the glow-arc transition.[5]

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4. Über rasche Vorgänge im Quecksilbervacuumbogen, O. Engelbrecht. *Archiv fuer Elektrotechnik*, Berlin, Germany, volume 36, 1942, page 515.

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O. K. Marti (Allis-Chalmers Manufacturing Company, Milwaukee, Wis.): In spite of the fact that during the last few years many different kinds of vacuum switches have been put on the market, the literature still does not give any theoretical explanations of the various phenomena which occur during the interruption of an arc in vacuum. I am sure, therefore, that many investigators in this field will appreciate the splendid and thorough work done by Doctor Koller and that they also will feel indebted to Professor L. B. Loeb of the University of California who fostered this research. I had the good fortune during my stay in California to be able to follow some of this work from time to time and also to see the switch in operation under different conditions.

When studying the present circuit breaker art, especially the means employed for arc interruptions, one is impressed very much by the ingenious methods and contraptions which are being used to increase the rate recovery of the dielectric strength of the medium in the arc path. This is chiefly caused by the fact that it is very difficult to fill the space between the parting contacts with the dielectric medium, be it oil or air, in a very efficient way, and this is especially difficult if the opening speed of the contacts is very high as is the case in all modern breakers. The only other way to accomplish this same thing was by pushing the arc out of the space traveled by the contact. It easily can be seen that in connection with the vacuum breaker this difficulty does not exist because the dielectric medium used for interrupting the arc is present, nothing has to be forced or injected between the contacts, nor are any means necessary to force the arc to travel. In addition it has to be realized that vacuum has a dielectric strength far greater than any other interrupting medium in use.

I am in full agreement with the author that since in connection with the development of sealed tube rectifiers the art of removing gases from tube walls and internal parts has been perfected, it should be much easier to develop a high capacity vacuum switch consisting of a glass or steel container with bushings or inlets as used with sealed tube rectifiers.

Some of the most important results obtained during these investigations are

1. Very good interruptions still could be relied upon at a vacuum of 10^{-4} millimeter of mercury, which is quite significant considering the fact that at a vacuum of 10^{-7} millimeter of mercury there are still about 10^{9} gas molecules per cubic centimeter present. From experience with rectifiers, especially sealed tube rectifiers, such vacuums can be obtained easily in glass or steel vessels if properly evacuated and heat treated.

2. The pumping action and getter action of this type of switch was such that the vacuum did not seem to decrease after hundreds of operations if the contacts and shields were designed properly in spite of the fact that some of the material is removed from the contact during each operation. Some of these favorable results have been obtained probably also because that part of the gases released from the anode and cathode spots have been absorbed by them after each single arcing episode.

3. The arrangement and shape of the anode and cathode contributes a great deal to the performance during an interruption.

4. The amount of material which is being removed from the cathode depends, as could be expected, on the type of contact material but is extremely small if the vacuum is maintained below 10^{-4} millimeter of mercury. It also should be noticed that the cathode and not the anode loses material.

However, it seems to me that the author made the most valuable contribution to the understanding of arc interruptions by his analysis of the mechanism of the arc on the basis of much data collected by means of special cathode-ray oscillograph and high speed camera setups which required very painstaking work. An excellent understanding of the arcing episodes can be obtained from a study of Figure 7. Following his theoretical outlines it easily can be imagined that the operation would differ materially from that of a high vacuum tube as there is no longer a clear path for the electrons, especially at the time of contact separation, because of many the emitted electrons will collide with gas molecules, a condition which will improve naturally after each successive episode.

At the end of the paper the author gives some of the essential design features which could be tested by future investigators. It would seem that a breaker built on his present layout could handle only currents up to 200 amperes per single contact. However, it has to be realized that these interrupting tests were made with direct current and it therefore can be assumed that a proper arrangement of the opening mechanism may permit the interruption of much higher current values in case of alternating current.

It is hoped that this investigation, which was carried out with somewhat limited facilities, will stimulate some further work along this line and I am sure further investigators will be glad to have, as a basis, Doctor Koller's results and the scheme for opening the contacts as well as the splendid optical method for measuring the speed of separation of the electrodes.

Co-ordination of Insulation and Spacing of Transmission Line Conductors

Discussion and author's closure of paper 46-175 by W. W. Lewis, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 690–4.

L. M. Robertson (Public Service Company of Colorado, Denver, Colo.): The electrical industry owes an expression of gratitude to Doctor Lewis for obtaining the data and presenting this paper. It will be a very valuable guide for transmission line designers and for a review of practice by operators who may wish to improve the operation of transmission lines.

There are a few factors which might be mentioned in connection with this subject in addition to those in the paper.

The experience of the Public Service Company of Colorado has indicated that contamination of insulators is a factor which must be considered in determining the size of pin type units or the number of disks in a string. Altitude also should be considered since an increase of altitude requires an increase of insulation on lines and stations.

It also has been determined that the use of wood in crossarms and crossarm braces may be of considerable value in line insulation and operation. This applies to lightning difficulties as well as those caused by birds. A 44-kv double-circuit line with pin type insulators on steel arms on steel poles and with overhead ground wire had a poor operating record. When the steel arms were replaced with wood arms and braces the troubles were eliminated.

A spacing between conductors and between conductors and ground wires may not be satisfactory in some cases where the terrain is hilly or mountainous and the winds blowing in such a direction and nature as to produce severe upcast of the conductors. Contact between ground wire and conductors has been obtained on a line of horizontal configuration where the phase spacing is 18 feet and the ground wire six feet above the plane of the conductors.

In general we have found that the data presented conform closely to that which has proved to be satisfactory in most cases in this territory.

M. S. Oldacre (Commonwealth Edison Company, Chicago, Ill.): The paper gives an excellent summary of transmission line practice and co-ordination of insulation that will serve as a useful starting point in transmission line design. It should be realized that the recommendations are based on the average of many averages. The actual design should be engineered completely with due consideration to the various limiting elements of line construction. This will allow for clearance as it is required and will give the most efficient use of material with the best operating results.

A comparison of recommendations in the paper with the dimensions of two existing lines may be helpful in considering how satisfactory the recommendations may be. The 230-kv line was built in 1940 and has had a very low outage rate because of lightning or dancing conductors. The other line, which was built in about 1928, has had a consideralby higher outage rate, and wherever possible alterations have been made to obtain better performance. The data are shown in Table I of this discussion.

The effect of ice or snow on the conductors is mentioned too briefly, and the effect during foggy weather of dirt accumulation on the insulators should be considered.

There seems to be a very large amount of literature on the subject of ice and snow on conductors and the frequently resulting dancing of conductors, but a critical study of it will produce no definite answer that will serve as a design basis.

If adequate allowance is not made for dancing conductors, there may be a loss of the lines at the time they are needed most or else some expensive mitigative measure may be required. Such a measure might be increased current loading of part of the lines while parallel lines are taken out of service, or might be the short-circuiting of a line and supplying it directly from the bus or from a separate transformer or generator depending on the line characteristics. Either one of these methods is an expensive one both from an investment and operating point of view.

The amount of clearance to allow for dancing conductors cannot be determined adequately from present data. The amplitude of the conductor movement may be nearly twice the sag of the conductor, if there is only one loop of dancing per span, but may have a maximum amplitude about equal to the sag if there are two loops. If there are more than two loops per span the amplitude will be reduced. The dancing conductor with one loop per span sometimes is designated as vibrating at its fundamental frequency and other times at the first harmonic. Therefore it is more definite as a practical matter to speak of the number of loops of vibration per span.

The conductors frequently dance in an elliptical orbit, but there are reports of conductors dancing vertically up and down. Thus the amplitude and direction of the dancing are important elements to consider in selecting conductor clearance.

It would be advantageous if anyone having helpful data on dancing conductors would make the information generally available by publication or correspondence between interested operating engineers.

Possibly further investigation would develop mitigative measures or devices that could be applied to the conductors to reduce dancing to an amount that will not cause serious trouble. No specific methods are offered but only the speculation that something might be developed.

The dirt and fog question is well worth more consideration than it has received. This trouble usually is concentrated in rather small areas and additional construction costs in such cases may be justified.

On some particular lines, the combination of dirt and fog is more of a menace than lightning. That is, there have been more outages caused by dirt and fog than by lightning.

It seems a hopeless and useless task to try to prevent all flashovers resulting from dirt and fog, but some remedies are available that have been satisfactory.

1. Cleaning the insulators by hand, either on the line or after replacement.

2. Washing the insulators with a high pressure water stream while the line is alive.

3. Providing additional insulation that may be sufficient to eliminate all dirt and fog flashovers or else eliminate them if the insulators are washed once a season just before the usual foggy weather.

There are good indications that the last method can be accomplished successfully with special insulators having increased creepage distance, particularly in surface that remains relatively dry. An alternate method is an increased number of insulator disks where suspension insulators are used. This may be relatively expensive if clearance to towers is maintained.

All of the foregoing text can be summarized by saying

1. The possibility of dancing conductors must be considered and allowed for where necessary.

2. The effect of dirt and fog must be considered and provided for.

3. Additional observation and investigation by those having trouble because of dancing conductors, dirt, and fog would be very valuable to the industry as a whole to obtain a better design basis for future lines or for possible alterations to existing lines.

Abe Tilles (Pacific Gas and Electric Company, San Francisco, Calif.): In applying the basic considerations of insulation co-ordination to the detailed selection and design of an insulator string and supporting structure a great variety of considerations need careful attention. When one tentatively has selected a design and wishes by way of check to make a general comparison to other lines, he is likely to find, as documented by the author, that a great variety of usages are in existence and that a limited comparison is confusing and may be very misleading. Accordingly, the comprehensive assemblage of data here presented and simplified is very welcome and useful as a general comparison and guide, as well as being very helpful as a rapid first cut in design in many instances.

Those who have labored in detail with the considerations involved will appreciate how skillfully and compactly the author has assembled and simplified, in gross outline, the various factors involved in the problem.

In a recent design of a major 230-kv line, because of the existing dust and weather conditions together with quite low footing resistance, wet flashover at 60 cycles was controlling at critical wind conditions. A somewhat different proportion of spacings than that listed by the author was selected accordingly. The loan by the author of a preliminary copy of his paper was of interest as a general check and was appreciated greatly.

Table I. Comparison of Recommended Practice and Actual Line Dimensions

	138-Kv Line		230-Kv Line	
	Recommendations in Paper	Data From Actual-Line	Recommendations in Paper	Data From Actual Line
Minimum number of insulator disks	7		12	
Maximum number of insulator disks	10		18	
Number of disks for average practice	9	12 (5-inch spacing)	15	18
60-cycle wet flashover, kilovolts	375	450	600	690
Impulse flashover, kilovolts	860	1,000	1,345	1,585
Clearances, feet				
Figure 5A			9.5	9.5 min.
Figure 5B			14.25	15.0
Figure 5C			19.0	28.8 to 37.5(d)
Figure 5D			28.5	33.3
Distance between ground wires, feet			28.5	46.3
Clearances, feet				
Figure 6A	5.0	6.75	9.5	10.4
Figure 6B	7.5	9.25 min.	14.25	15.+
Figure 6C (conductor to conductor)	10.0	15.0	19.0	20.5 to 26.0†
Figure 6C (ground wire to top conductor)	10.0	12.0	19.0	23.0 to 39.0‡
Figure 6E	2.5 to 7.5	5.0* 4.5** 8.0†	4.75 to 14.25	4.0
Location of ground wires	Over outer conductor		Over outer conductor	Over outer conductor

* Top conductor inset from center conductor.

** Bottom conductor inset from center conductor.

† Ground wire inset from center conductor.

‡ Dimension *C* is larger for the longer spans.

H. A. Frey (Locke Insulator Corporation, Baltimore, Md.): Doctor Lewis has presented a very interesting and practical rationalization of transmission line insulation. It should be noted that his insulation levels are established principally by the balance between the voltage expected from switching surges and the wet flashover values of the insulators. The user thus is left free to select a particular insulation level corresponding to the magnitude of switching surges which may be probable on the system in question. In so doing it must be borne in mind that the rated wet flashover values of insulators result from an arbitrary yard stick. They are not necessarily the lowest voltages at which the insulators will flashover. The rated wet flashover voltage is merely the average value at which flashover will occur when tested in accordance with the standards of the AIEE.[1] In service under varying condition of precipitation and with the insulator surface fouled with dirt to varying degrees, it is quite possible that lower wet flashover voltages may be obtained. Under some conditions it therefore may be desirable to use more conservative insulation levels.

REFERENCE

1. INSULATOR TESTS. American Standard C29.1. 1944.

W. W. Lewis: One purpose of the paper is to give a means for quickly estimating the minimum insulation that should be used on a transmission line. In practice this insulation may be increased somewhat to allow for local conditions such as excessive lightning, high footing resistance, dirt and fog. Finally when the number of disks is selected, all spacings to grounded members of the tower between conductors and to overhead ground wires should be co-ordinated in order to prevent a weak spot for flashover and prevent waste of insulation or spacing. The paper gives a method of readily selecting the co-ordinated spacings.

In selecting the insulation, switching surges and wet flashover have been given the main consideration. However, as brought out by all of the discussers, the effect of dirt, dust and other contamination during foggy or wet conditions may lower the flashover of the insulators considerably and require more conservative insulation levels, special insulators with large creepage distances, or a program of periodic washing.

M. S. Oldacre stresses the phenomenon of dancing conductors and mentions the large amount of literature on the subject which does not produce a definite answer that will serve as a design basis. This phenomenon seems to be localized in a belt adjacent to the Great Lakes. The experience of companies operating in this region is teaching the designers how to lay out their lines to cope with this trouble. Companies in a somewhat wider and longer belt experience trouble caused by the accumulation of snow, ice, or sleet on conductors, not usually accompanied by dancing. They also have learned by experience how to overcome this trouble and to continue in operation. As Oldacre states, an actual design should be engineered with due consideration to all of the local limiting features.

The author wishes to thank each one who so kindly has offered a discussion of the paper.

Current and Torque of D-C Machines on Short Circuit

Discussion and author's closure of paper 46-136 by T. M. Linville, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 956–65.

B. Litman (Westinghouse Electric Corporation, East Pittsburgh, Pa.): T. M. Linville's paper does much to systematize the calculation of short-circuit currents in d-c machines and introduces the previously neglected factor of the a-c reactance voltage which opposes current flow. Apparently this factor has been overlooked in the past because on high voltage machines it exerts only a minor influence on the magnitude of the short-circuit current. In order to check the actual magnitude of this voltage, short-circuit oscillograms have been taken on a comparatively low voltage generator which had probe coils wound about the main pole tips to determine the actual loss in flux. The values obtained, which are listed in the following text, indicate fairly good agreement if a correction is made for saturation of the teeth in the commutating zone.

In equation 27 of Appendix I application is made of the theorem of constant flux linkages, and the field leakage coefficient n_f is introduced. It appears that this should be the over-all effective leakage coefficient introduced by G. E. Frost[1] in order to include the effect of armature leakage flux which reduces the air gap flux and yet does not link the main field winding.

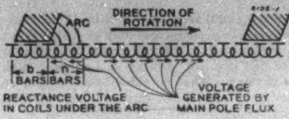

ARC
DIRECTION OF ROTATION
b BARS
n BARS
REACTANCE VOLTAGE IN COILS UNDER THE ARC
VOLTAGE GENERATED BY MAIN POLE FLUX

Figure 1

We wish to introduce a somewhat different explanation for the equivalent contact drop caused by a-c reactance voltage. The differential equation for a coil undergoing commutation (equation 59 of Appendix IV) is solved to give

$$i_s = A\epsilon^{pt}$$

and the constant A is calculated by assuming that the total current adds up to the line current per arm. This tacitly assumes that the brush contact drop is large enough to accomplish this successful commutation, and that if the reactance voltage were twice as large, the contact resistance drop somehow would be twice as large so that commutation still would be successful. Actually, very little current will be collected through the brush and most of the current will appear in an arc at the trailing edge.

Figure 1 of this discussion shows such an arc at the trailing edge extending effectively over n bars. The current in these coils thus is changing from peak value in one direction to peak value in the other direction over a span of n bars, whereas if normal commutation existed this change could take place under the brush in a span of b bars. These coils therefore have an average reactance voltage of $E_R \times b/n$ per coil, and as the figure indicates there are n of these coils in series with their voltages opposing the generated voltage. Therefore the generated voltage should be reduced by a total amount $E_R \times b$ and the equivalent brush contact resistance is E_Rb/I ohms which is the same answer as given in equation 55, Appendix IV of Linville's paper.

The calculation of an equivalent resistance from the value of E_R at rated load assumes that this reactance voltage changes linearly with load current. Actually the slot leakage fluxes which give rise to the major portion of this reactance voltage will saturate the teeth before short-circuit current is reached. On the machine tested, this saturation is reached at less than half the peak short-circuit current so that the equivalent brush contact resistance is only a little more than half the value which otherwise would be expected. The values obtained on test are

Machine rating = 1,500 kw, 250 volts, 5,000 amperes, 514 rpm
Flux before short circuit = 1.0 = 250 volts
Flux loss during short circuit = 0.511 = 128 volts
Net flux after short circuit = 0.489 = 122 volts
Calculated equivalent brush contact drop caused by a-c reactance voltage (neglecting saturation) at 10.2 times rated current, which is the actual short-circuit current obtained = 103 volts

In obtaining the latter value, the magnitude of E_R was obtained from equation 63 of Appendix V.

Since this voltage is almost as large as the net generated voltage after short circuit, it is evident that a correction for saturation is necessary. In making this correction we have assumed that the teeth are equivalent to air paths for flux densities above the saturation value of the iron. The value then obtained for this voltage drop is 54 volts.

Net voltage available = 122 − 54 = 68 volts
Armature circuit resistance = 0.00113 ohms
Peak short-circuit current = 68/0.00113 = 60,000 amperes

The actual value of current obtained was 61,000 amperes.

REFERENCE

1. The Short-Circuit Characteristics of D-C Generators, G. E. Frost. AIEE Transactions, volume 65, 1946, June section, pages 394–402.

T. M. Linville: B. Litman's comments are interesting and very much appreciated. He disagrees with the paper in saying that the a-c reactance voltage exerts only a minor influence on the magnitude of the short-circuit current in high voltage machines. The analysis made in the paper and the author's tests make it evident that the a-c reactance voltage actually is a major factor opposing the current flow and generally the largest factor in all machines, including high voltage ones.

This is made apparent by recognizing that with linear commutation at ten times normal current the reactance voltage would be ten times as much as at rated current. Then consider that at the instant of this peak current the commutating pole flux has not had time to rise and the nonlinear commutation shown in Figure 3 exists. Then one has to think of reactance voltage in terms of 50 times normal. As a matter of fact the multiplying factor on account of nonlinear commutation is equal to the brush overlap in segments according to equation 55 which Litman substantiates. Generally the brush overlap on high voltage machines is large so that the effect of a-c reactance voltage opposing the current flow on these machines is large.

Litman suggests that the effective over-all leakage coefficient given in Frost's paper should be used in Equation 27 in order to include the effect of armature leakage flux on the air gap flux. The method of the paper in effect develops Frost's theory to a point where calculations from machine dimensions can be made. This is done by determining separately the air gap flux (or pole to armature flux) and the flux across the space from pole to pole. The former is a function of armature current and field current and is determined by equation 23. To derive this equation the field magnetomotive force and armature magnetomotive force are combined and multiplied by the magnetic permeance. In effect this product is integrated around the air gap taking into account the varying values of magnetomotive force and permeance. This then determines the net air gap flux. The pole-to-pole flux is determined by the second term of equation 27 considering it proportional to the field current.

Litman is believed to be saying that all of the net air gap flux between pole and armature, equation 23, does not link the field winding as assumed in equation 27.

This is the same as saying that all of the flux going directly from pole to pole does not link the field winding.

For these statements to be true there would have to exist in the pole iron a molecular magnetomotive force which acts with the same polarity as the field magnetomotive force and becomes effective as the iron of the pole shoe saturates. This is improbable because the over-all change in the magnetic permeance of the pole shoe caused by saturation acts to oppose the field magnetomotive force always. Furthermore, if equation 27 does not truly represent the field linkages because some part of the air gap flux ϕ_d goes directly from pole to pole without linking the field, then that portion of the pole-to-pole flux which does link the field must be greater than what is accounted for in equation 27. Such an error in the method would make the transient field current calculated by the paper smaller than actual. Therefore the justification of the paper lies not only in its logic but also in comparison of the calculated field current with test results. Figures 1, 4, and 5 show a good check substantiating the method of the paper.

It is considered that there is no appreciable change in flux in the commutating pole circuit, affecting the armature electromotive force, in the time interval between the instant of short circuit and the moment of peak current.

It seems axiomatic that the brush contact drop, or arc drop, from brush arm to commutator always at least must equal the reactance voltage caused by current reversal in the armature coils when the flux of the commutating pole is negligibly small. Otherwise current reversal would not occur. This equivalent brush contact drop, or arc drop, is calculated in the paper and when divided by the peak short-circuit current is equal to the equivalent resistance r_b. Litman is to be thanked for demonstrating that the expression for r_b, equation 55, is independent of the number n of bars over which the arc between brushes and commutator extends. The paper states this fact without demonstrating it.

Litman's figures for the machine he tested cause him to conclude that the value of r_b should be corrected for saturation of the armature teeth. Otherwise the set of figures given are not consistent. The author does not believe that this is true. The armature teeth underneath a commutating pole at the instant of peak short-circuit current carry substantially only the flux crossing the armature slots caused by magnetomotive force in the slots. It is the reversal of this flux which causes the reactance voltage E_R, which determines r_b. In some machines saturation is present in the teeth at the beginning and end of this reversal period but in between the reactance voltage reaches the full unsaturated value. In general this saturation has small influence and no correction should be made. The results of the paper substantiate this conclusion.

It seems that the value for r_b obtained by Litman using the method of the paper is more nearly the correct value. The incompatibility in his set of figures evidently is caused by the value used for the flux loss during short circuit which is the measured loss during short circuit but not at the immediate instant of peak armature current. The paper is believed to be correct in assuming that the maximum current is reached so quickly that there is negligible loss in the flux linkages of the field circuit in this time interval, that is, the theorem of constant flux linkages can be applied. If so, the immediate loss of flux cannot be 0.511 because to hold constant flux linkages in the field this loss of air gap flux must be balanced by at least an equal gain of pole-to-pole flux. As the pole-to-pole flux of the machine with normal excitation should not be more than about ten per cent of normal exciting flux and the permeance of the pole-to-pole path is fairly constant until after peak armature current is reached, it is apparent that the field current must rise to more than six times normal to hold constant linkages. So great a rise in field current never has been observed to the author's knowledge. However, sometime after maximum current when the induced field current has diminished, the loss of air gap flux and field flux linkages will be as great as 0.511. It should be re-emphasized that r_b is derived to represent the effective brush arm to commutator resistance at the instant of maximum short-circuit current (maximum current is the important calculation to be made), and thereafter it will be too large a quantity. The following set of figures are possibly correct for Litman's machine at the moment of peak current:

Exciting flux before short circuit = 1.0 = 250 volts
Exciting flux loss at moment of maximum short-circuit current = 0.312 = 78 volts
Net flux at moment of maximum current = 0.688 = 175 volts
Calculated brush arm to commutator drop = 103 volts
Net voltage available = 172 − 103 = 69 volts
Armature circuit resistance = 0.00113 ohm
Peak short-circuit current = $\dfrac{69}{0.00113}$ = 61,000 amperes
Field leakage flux linkages = 0.1 times normal exciting flux linkages
Initial field linkages = 1.1
Leakage linkages at maximum current = 1.1 − 0.688 = 0.422 ampere
Maximum transient field current = 0.422/0.1 = 4.22 amperes

Four times normal field current is much more consistent with tests made by or reported to the author.

Semiconducting Shielding for A-C Power Cable

Discussion and author's closure of paper 46-118 by Edwin J. Merrell, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 605–09.

R. W. Atkinson (General Cable Corporation, Bayonne, N. J.): The data given by Edwin J. Merrell should be very convenient for use in determining conditions resulting when there is a considerable voltage drop along the length of a semiconducting shield. I am inclined to feel, however, that the approach generally should be along another direction, namely that the total voltage drop should never be permitted to be more than a relatively small fraction (say ten per cent) of the full voltage of the cable. When this is the case no significant error is introduced by calculating the current in the shield as though it were metallic. In this simple calculation any slight error is on the safe side and is insignificant in any event in terms of the engineering data.

There seems no real reason to use a shielding tape and then let it only partly perform its function by permitting a voltage to ground that is a large fraction of what would exist without the shield. Moreover, there seems no excuse for permitting a loss in the region of the grounding contacts sufficient to produce even a slightly significant effect on the temperature rise and available carrying capacity. These points are emphasized when it is remembered, as the author points out, that the loss and the voltage drop are proportional to the square of the spacing of the ground points and thus that these are under rather simple control. This view seems not to be in conflict with that of the author except for emphasis.

E. W. Greenfield (Anaconda Wire and Cable Company, Hastings-on-Hudson, N. Y.): It is gratifying to see this paper finally published. More than five years have passed since the author first indicated his work on semiconducting jackets at the National Research Council meeting in Williamsburg, Va.

Independently, the writer had come to the conclusion that voltage gradients along a semiconductive jacket could be interpreted best on the basis of distributive infinitesimal parameters. In our work on nonmetallic conductive jackets, we have used transmission-line differential equations which are similar in most respects to those shown by the author. However, we do not form an artificial term of RL^2 as parameter but instead we use the conventional terms for the cable, namely propagation constant and characteristic impedance.

To present a physical picture of the underlying phenomena giving rise to the use of transmission-line theory in this work, we offer the following.

Any conductor maintained at an alternating difference of potential with respect to earth will carry leakage and charging currents in addition to load current. The former are functions of the quality of the insulation and the geometric relation of conductor to earth, that is, shunt conductance and capacitance. Being distributive in nature, these shunt currents (measured in the conductor) are also proportional to the length of conductor.

For a shielded cable, that is one having an outer conductive jacket, the leakage and charging currents flow longitudinally along the central conductor while at the same time they pass proportionate amounts, as required, radially through the insulation. The conductive sheath then collects substantially all the radially passing current and carries it back longitudinally to the energy source or to a grounding point.

For a nonshielded cable exactly the same sequence of phenomena takes place. In general the total amount of radially passing current will be less since the shunt conductance and capacitance to earth are smaller. However, the requisite total amount of leakage and charging current will again flow longitudinally along the conductor, gradually dissipating radially through the insulation as demanded by the per-unit shunt parameters. The current flow through the insulation thus appears as a distributed charge on the surface of the insulation. As a result of this charge a longitudinal current in general will flow along the insulation surface between grounded points or between regions of low resistance contact to earth. For a given surface charge the amount of longitudinal surface current that flows then will depend on the longitudinal resistance of the cable surface and the distance between grounded points. That is, more resistance, less current; more length, more charge (but also more resistance). Hence current can both increase or decrease with increase of length depending on other factors.

The existence of charges and (or) current flow at the insulation surface gives rise to a potential along the cable covering. The voltage to ground at any point along an unshielded insulated cable thus depends on all the conditions previously outlined for current flow. For an infinitely long unshielded cable (infinite distance between earth contacts) the surface voltage to ground is simply proportional to the ratio of the insulation capacitance to the earth capacitance.

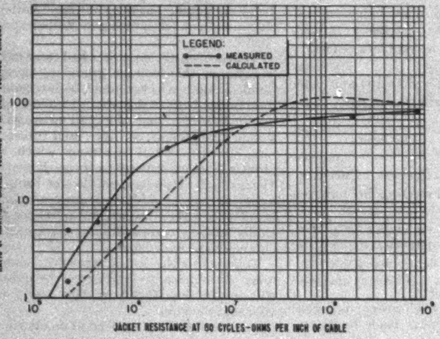


Figure 1. Calculated versus measured surface voltages for a length of 1-conductor 500,000-circular-mil 6/64-inch rubber-insulated cable

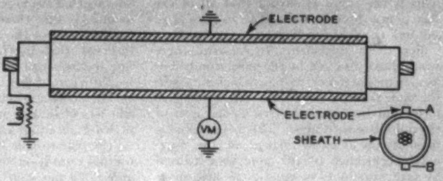
Ground ring spacing is six inches

For finite lengths and surface resistances the voltage to earth is more complex and is determined by all the factors discussed.

To verify from another laboratory the experimental data given by Merrell, we present the two curves of Figure 1 of this discussion showing the voltage rise along the jacket in per cent of applied voltage for a length of 1-conductor 500,000-circular-mil 6/64-inch rubber-insulated cable. Ground ring spacing was held at six inches and the jacket a-c resistance varied over a range from 0.2 to 600 megohms per inch of cable length.

The solid curve shows the data taken experimentally using a special (0.65 micromicrofarad) electrostatic voltmeter. The dotted curve represents calculated data using transmission-line differential equations. The agreement is seen to be excellent at high values of jacket resistance. At low values the experimental data are somewhat erratic but the shapes of the two curves indicate general agreement.

Figure 2



W. T. Peirce (American Steel and Wire Company, Worcester, Mass.): Edwin J. Merrell's paper is of value in setting one boundary point for shields that have far more resistance than conductance. It mainly considers the case where the natural conductance of fibrous materials containing moisture or lead base pigments has been found to exhibit some shielding effect and thus aid in reducing stresses from the surface of the insulation to occasional grounds.

The greatest application of nonmetallic shields in present day cables concerns rubberlike materials with a designed resistance in the order of 10,000 ohms per foot of cable used as a supplement to metallic shielding or in conjunction with ground wires. Under such conditions the shield is maintained at substantially ground potential and the principal terms of Merrell's formula lose significance. The shield is less than one-half of one per cent of ground potential at any point and its function merely is to counteract the inevitable looseness that develops between the metal shield and the insulation caused by mechanical or thermal changes in the components of the cable.

In the worst case all the shield has to carry is the circumferential charging current from the back of the conductor to the ground wire in longitudinal contact with the shield. Figure 2 and Table I of this discussion, show the potential so developed with a wide range of shield resistance.

Merrell has considered only the single phase condition whereas by far the more common use is for the three conductors of the circuit to be associated together closely. In such case the 3-phase charging currents balance out per-unit length of cable and no longitudinal flow of current is involved. Here, even in the absence of ground wires or other metallic shield, only the circumferential resistance to the point of contact of the three conductors is involved, and the adequacy of the shield can be established without resort to complex mathematical treatment.

The great value of these nonmetallic shields is to permit the cable designer to be certain that corona has been eliminated from both outer and inner surface of the insulation and thus utilize insulation that has the best mechanical or electrical properties without the necessity of compromising in order also to obtain corona resistance.

Perry H. Ware (Simplex Wire and Cable Company, Cambridge, Mass.): Application of the plots in Edwin J. Merrell's interesting and timely paper to several practical cases of cable design have brought out the following suggestions for further saving of time and computation.

The voltage vector diagram of Figure 3 in the paper presents the following difficulties in its use:

1. Capacitance is given for only a few discrete values.
2. Evaluating the voltage drop along the shield and across the dielectric involves scaling the lengths of the vectors on the plot.
3. The accuracy definitely is limited for small values of voltage drop along the shield.

In many cases it is the magnitude of the voltage drops which is significant for practical uses; the phase angle is relatively unimportant. On this basis Figure 3 was designed to give the magnitudes of the voltage ratios for 60-cycle a-c line voltage. The unit of the abscissa CRL^2 was chosen so as to allow a continuously variable range of capacitance as well as of resistance and ground spacing. An extension of Merrell's mathematics further shows that for values of CRL^2 below 10^2

$$V_D/V_L = 1 \tag{1}$$

$$V_B/V_L = 0.000188CRL^2 \tag{2}$$

R. C. Waldron (The Okonite Company, Passaic, N. J.): Edwin J. Merrell has presented an interesting method of measuring the a-c characteristics of semiconducting coverings. The descriptions given in "Measuring Technique" are not quite clear and it would be helpful if he would include a wiring diagram showing the various connections to the Schering bridge. From this description it is not clear whether the loops are connected directly to the bridge with one loop being connected to a microammeter

Table I

Size	Insulator Wall Thickness, Inch	Charging Current, Microamperes Per Foot Per Kilovolt	Sheath Resistivity		Volts Difference A to B Per Kilovolt Applied
			Ohms Per Cubic Inch	Ohms Per Foot	
6	15/64	18.5	530	65,000	0.014
6	15/64	18.5	160	11,000	0.007
6	15/64	18.5	850	125,000	0.302
6	15/64	18.5	1,500	340,000	0.471
4/0	11/64	70	250	19,000	0.028
4/0	11/64	54	250	33,000	0.022
500,000	16/64	55	650	300,000	0.064

at the same time or whether these are separate measurements. If these are connected simultaneously, it would appear that some of the current may be shunted by the bridge and not pass through the microammeter.

Work which we have done in connection with the study of discharge from braided cables had indicated that a large part of the charging current can be drained from a 4-foot sample having a wet braid by grounding in the center. For example, at 2,000 volts alternating current, the measured current from a 53-inch length was 331 microamperes while the theoretical current should be 442 microamperes. Working back from the measured current by a method similar to Merrell's we obtain a surface resistance of 0.78 megohm per inch. The d-c resistance measured about 60 megohms per inch. We would like to know whether the author has any comparative values of a-c versus d-c resistance that show this magnitude of difference.

Similar a-c measurements on an okoprene covered cable show an a-c resistance of 200 megohms per inch against a d-c resistance of 200,000 megohms per inch.

Experience has shown that braided cables are not satisfactory where drainage of any appreciable length can occur. Okoprene cables, however, are satisfactory. This would indicate that 0.78 megohm per inch is too low for a nonshielded cable and that 200 megohms per inch is satisfactory. The limiting value is probably somewhere between these points.

For Merrell's illustration he shows a cable operating in cable rings. He points out that in this example with a surface resistance of two megohms per inch there would be a 12-degree rise in temperature at the point of current discharge. This assumes uniform current flow around the entire circumference of the cable, whereas with the cable in rings we would expect almost point contact with the rings which would result in much higher concentration of current at this point with a correspondingly higher temperature. In this case also he obtains a voltage in the center between rings of 5,500 volts to ground which would be approximately 10,000 volts between phases if three conductors were installed in the same rings. This voltage would cause sparking at the center in case of close proximity or occasional contact between conductors, or between conductors or some grounded object such as tree limbs. This example has the same order of magnitude for the surface resistance as braided cable which is known to be unsatisfactory. I am wondering if the value of resistance for the covering chosen in this example is considered satisfactory and whether any general recommendations can be drawn from their experience as to the maximum values of resistance which are satisfactory.

In our design of semiconducting coverings we have worked to much lower values and have assumed that the cable covering at no point should reach a voltage high enough to cause shock to personnel touching the cable nor to cause sparking in case of occasional contact with adjacent conductors or grounds.

T. F. Peterson (American Steel and Wire Company, Cleveland, Ohio): May I express my commendation and appreciation for the fine rigorous mathematical treatment of semiconducting shielding put forth by Edwin J. Merrell.

Some 20 years ago I had occasion to derive the equations for the thermal resistance from the copper conductors of type *H* (Hochstadter copper tape shielded) cable to the lead sheath. I was happily surprised to find, as Merrell has found, that these second order differential equations were almost identical to those for long distance transmission lines. Frequently the greatest usefulness of such studies is derived from their application to "boundary conditions" of the systems and I would like to direct attention to them.

I would like also to comment on the historical references included in the paper.

Merrell hails Lowenberg as an originator of semiconducting shielding and makes no reference to the vast number of patents and comprehensive literature on this same subject since 1928. Careful reading of the reference will reveal that he was concerned with arcing between weatherproof braids and cable installed in fiber ducts. Nowhere does he use the terms shielding and semiconducting, nor is there any evidence that the test samples which were laid on a concrete floor were either grounded or shielded. Apparently very high voltages existed on the braids of all test samples, but under certain circumstances (perhaps highly resistive concrete or widely removed grounds without electrical connection) arcs and discharges occurred only at high conductor voltages. Lowenberg indicates that braid burning caused by arcs was attributed by some to the drainage of charging current along damp braids to points where the cable was in contact with the duct. Others reasoned that the braids were dry and that dielectric relationships determined stress distribution which was such as to cause corona discharge in certain air gaps. Anyone familiar with these problems should have known at that time that both theories may have accounted in part for the observations since the whole system represents a complex combination of parallel and series resistances and capacities from conductor to ground.

In view of these uncertainties, it is not surprising to find that Lowenberg observed visible discharge at 4,000 volts in the case of weatherproof braided cable and discharge beginning at 9,500 volts in the case of flameproof (presumably asbestos) braid. In his conclusions he seems to ignore the second set of tests in which the weatherproof braided cable showed discharge beginning at 10,000 volts whereas a second portion of the same cable with a flameproof finish showed "pronounced discharge" at 4,000 volts. Apparently anything could happen and almost any voltage could appear on the various braided cables, and perhaps the only reason the flameproof braided cable behaved better in service was because of its resistance to arcs and not because of partial shielding.

For more than 20 years I have done research on various types of shielded cable and for more than 18 years have been engaged in the development of what I have called semiconducting shielded cables. I cannot accept the author's definition of such shielding when he defines it as "the practice of partially confining the electric field within the insulation by means of semiconducting covering." Such a cable would not be shielded in the sense that it would be safe to handle, prevent corona discharge, or prevent interference from or with the insulated conductor. Merrell's cable would be

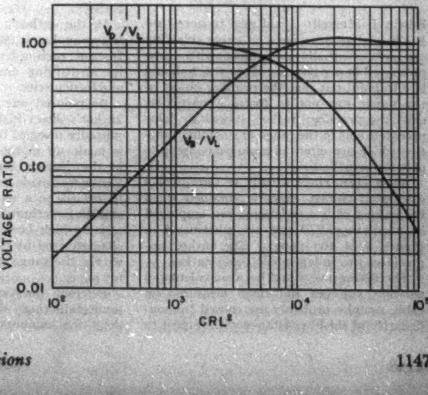


Figure 3. Voltage ratios for shield and dielectric of single conductor cable versus CRL^2

semishielded and I think he will agree that by a large margin the range of his studies covers such types of shielding. (It should be noted, however, that unlike Lowenberg very positive grounds are applied to certain portions of the semiconducting layers.)

These points are stressed. I contend that semiconducting shielded cable should

1. Be shielded.
2. Confine practically all electric field within the insulation.
3. Prevent disturbances and interference.

In the practical embodiments of these forms of shielding which I have made over the past 20 years, various values of R and L have been employed but in every instance the product of RL^2 has been much less than ten. For example, in using semiconducting shielding with surface resistivity as high as 100 megohms per centimeter square (as cited in United States Patent 2,322,702) and with ground points separated by 1 inch (L equals one-half inch), RL^2 equals eight. In certain forms of aerial cable supported by metal rings from messengers, even if we take resistance as high as 10,000 ohms per foot with grounding points every 20 inches (that is L equals ten inches), RL^2 equals 0.1. It will be observed that when RL^2 equals eight, cosh 3.03 equals 1.045. When RL^2 equals 0.1, cosh 0.033 equals 1.00045. Whether by substitution in Merrell's equations or by reference to Figures 4, 5, and 6, one must observe that with these values of resistance the cable is shielded fully and the electric field is confined fully to the insulation. It also will be observed that the watt loss in the shield is zero and so there are no further questions as to carrying capacity or change in characteristics of the shield.

These comments are made not to depreciate the work of Merrell but to utilize it practically to assure engineers that there are no hazards either from voltage or heat when conventional semiconducting layers are employed in conjunction with proper grounding. This discussion, while primarily directed to 60-cycle power conductors, is equally applicable to such cables as high voltage radar pulse cable and fortunately Merrell's equations are broad enough to cover a range of frequencies.

In closing I would like to suggest that the author describe more fully the types of semiconducting coverings referred to in his paper.

Edwin J. Merrell: I am glad to note that R. W. Atkinson's views on cable shielding are not essentially in conflict with those presented in my paper. However, I should like to point out that the primary object of the paper was not to foster a particular shielding practice; rather, it was to point up extensions in the range of shielding practice which are often advantageous, in spite of being unorthodox.

An excellent check with E. W. Greenfield's calculated voltage curve is obtained when his values of a-c resistance are employed in calculating RL^2 and are substituted in Figure 3 of the paper. The author appreciates this independent confirmation.

The divergence noted between calculated per-cent voltages and those measured on cable samples probably are caused by nonlinearity of shield resistance with respect to current density. This certainly would be suspected if a different order of current flow occurred during the shield resistance measurements from that characteristic during the cable voltage determinations.

I am in substantial agreement with W. T. Peirce's remarks except that I should like to add that the hazard to personnel cannot be overlooked when such high conductivity semiconducting shields are employed. Adequate grounding must be guaranteed on all cable installations of this type.

Perry H. Ware is correct in pointing out the limits of the graphs in the paper. However, according to our experience, the accuracy with which semiconducting shielding properties may be predicted is of no higher degree than that attainable through rough visual interpolation with the graphs as presented. This is because of the nonlinearity of the electrical characteristics of semiconducting shielding materials.

When the shielding circuit has been characterized by a CRL^2 below 10^2, that is an RL^2 below two for C equal to 50 and below 100 for C equal to one, the author has treated it as a conducting shield. The solution may be obtained accurately and quite easily by assuming a simple T equivalent.

R. C. Waldron questioned the bridge and and meter connections described under "Measuring Technique." The microammeter is connected at the same time and is in series with the portion of the cable sample which it is discharging and the bridge. The parallel short-circuiting of the microammeter is of no consequence since this short circuit is in series with a section of the semiconducting shield which has a resistance of an order thousands of times greater than that of the ammeter.

The author has noted differences between the a-c resistances of semiconducting shields and the d-c resistance as great and greater than those reported by Waldron. In fact it was this phenomenon which caused the derivation of the a-c theory presented in this paper, so that experimental data could be employed effectively in the design of semiconducting shielded cable for 60-cycle operation.

The cable in the illustrative example was chosen primarily to demonstrate the use of the graphs. However, it is a feasible construction for many single phase circuits. It would not be practical at this voltage for a 3-phase self-supporting aerial cable unless the shield were flameproof and unless considerable radio interference could be tolerated.

In the author's opinion there is no optimum value or limit of shield resistance. Instead, each order of shield resistance has its identifying characteristics, its merits, and its demerits.

As pointed out by T. F. Peterson, the author's direct historical references are admittedly meager, but the lack in this paper is made up in the Hickernell paper noted in the introductory paragraphs. The selection was made to high-light what appeared to me to be a natural difference in cable shielding technique. Moreover, the materials which Lowenberg found efficient are characterized by electric conductivities well within the range covered by the theory of my paper.

Peterson questions the use of the term semiconducting shielding. Actually, my definition amounts to a shortening of that adopted by the Insulated Power Cable Engineers Association in 1942. The full definition, which follows, is in printed form.[1]

> "A semi-conducting shield for electric power cable is a shield which has an impedance such that without fully grounding the surface of the insulation it nevertheless performs the following two functions either singly or simultaneously:
>
> "(*a*). Provides protection for cable insulation by restricting longitudinal and tangential voltage stress on the cable surface adjacent to miscellaneous ground points to a gradient below that necessary to produce corona.
>
> "(*b*). Provides a measure of protection for human life by limiting the magnitude of cable discharge current at miscellaneous ground points."

According to Peterson's concept, materials suitable for use as semiconducting shields are limited to a small number since he employs a narrow range of resistances just below the point at which losses in the shield become apparent. The definition of the Insulated Power Cable Engineers Association is much broader in scope. It checks with the theory presented in the paper, which shows that many materials characterized by resistivities in the megohm range are sufficiently conducting to prevent corona discharge in particular cable installations by restricting unit voltage stresses tangentially on the cable surface to a low enough value.

REFERENCE

1. GENERAL SPECIFICATIONS FOR WIRE AND CABLE WITH RUBBER AND RUBBERLIKE INSULATIONS (first edition). Insulated Power Cable Engineers Association, section 6.3.

Modern Excitation Systems for Large Synchronous Machines

Discussion and authors' closure of paper 46-152 by J. B. McClure, S. I. Whittlesey, and M. E. Hartman, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 939–45.

J. H. Kinghorn and **F. M. Porter** (American Gas and Electric Service Corporation New York, N. Y.): The authors are to be commended on their excellent summary of modern excitation practice. To their list of types of excitation systems also might be added the main exciter with electronic pilot exciter and regulator. This consists of a small electronic exciter supplying the field of the main exciter and controlled by an electronic regulator and has given very satisfactory performance. If it receives its power from the machine terminals it, like the electronic exciter, must have a normal ceiling sufficient so that it can provide adequate excitation under the most severe external short circuit to which the system can be exposed.

From a performance standpoint there are three important characteristics to be considered in selecting voltage regulators.

1. Reliability.
2. Accuracy.
3. Speed of response.

All of the types of voltage regulators outlined by the authors are in use on the systems of the American Gas and Electric Company, and experience has shown that all meet the requirements of reliability and speedy response. As far as accuracy is concerned the impedance and electronic types of regulators are much better than the others. The direct acting type is fairly accurate, but friction and lost mechanical motion cause a certain amount of deviation which, however, can be overcome to a considerable extent if sufficient amplification is put into the actuating circuit. The indirect acting regulator has a dead band which limits its accuracy to the width of the dead band.

The advantages of accurate voltage regulation make it desirable to employ one of the types of continuously regulating regulators in all cases. For synchronous condensers located at load centers these advantages from the standpoint of service to customers are obvious. For generators an added advantage is the increased ease of parallel operation without the necessity of introducing excessive amounts of droop. For small synchronous condensers where less expensive regulating equipment is required the direct acting regulator has proved satisfactory when a small booster voltage is added to provide more speedy response at low excitation. For larger machines the impedance or electronic regulator is applicable. Two 32,500-kva turbine generators now are being installed which employ static type regulators with rotating amplifiers and, based on seven years of highly satisfactory operation by one such arrangement used with a water wheel generator, successful performance is confidently expected.

C. E. Valentine (Westinghouse Electric Corporation, East Pittsburgh, Pa.): Four types of voltage regulators are described in this paper, namely, direct acting rheostatic, indirect acting rheostatic, impedance, and electronic. The direct acting rheostatic regulator with modern constructional features is probably the simplest regulator mechanically and electrically and the easiest to understand of any modern type. It is an electromechanical type and as such has moving parts, but when this regulator is applied properly, long life and freedom from maintenance or replacement troubles has been indicated by many years experience. As our interest becomes aroused in new and different types of voltage regulators, we should not lose sight of the availability and practical usefulness of the direct acting rheostatic regulators for small and medium size generators. Improvements in this type of regulator can be expected.

The indirect acting rheostatic type regulator has been used widely for industrial, public utility, and central station generators and synchronous condensers. It has been used successfully on systems where the load is relatively steady and also where it is fluctuating. Many applications have been made using a series-type main-exciter field rheostat. When this type of rheostat is used, only one set of single pole high speed contactors is required to provide the desired rapid response of the excitation system.

With regard to the impedance type regulator, a considerable number of this type, using the intersecting impedance principle, were manufactured and used with small a-c generators about ten years ago. At that time these regulators were relatively bulky on account of the space requirements of copper oxide rectifiers of which more were available than selenium. Improvements in the art and better materials now have made it possible to build a better regulator of this type for use with special exciters of the rotating amplifier type to control the voltage of large synchronous machines.

An electronic voltage regulator is now available using in the voltage responsive circuit a filament controlled high vacuum tube. This method of voltage measuring makes it possible to feed the a-c generator voltage directly into the regulator without the use of rectifiers and filtering. A direct voltage proportional to the rms value of the alternating voltage can be obtained from the loading resistor into which the tube feeds. This is important, for example, where unusual wave form or high frequency alternating current is regulated. This type of regulator can be used conveniently to control an electronic exciter or a rotating amplifier exciter.

The authors discuss several special control features including upper and lower excitation limits. It is indicated that these special control features are introduced more readily into static regulator control circuits such as the electronic type. It should be noted however that under excitation and overexcitation limits have been provided and used successfully with indirect acting rheostatic regulators in a number of applications. One type of lower excitation limit used with generators has been an electromechanical element wherein a-c machine main field current has been balanced against a-c watt output. This variable underexcitation limit has been used to prevent the voltage responsive and regulating means from lowering the excitation beyond a desirable point. For overcurrent limit on synchronous condensers an a-c electromechanical current element has been used, thus automatically controlling the upper limit of excitation in place of sounding an alarm in case a predetermined load current is exceeded.

An item of interest to many operators is the modernization of old excitation systems. In such cases it is important to determine accurately the characteristics of the existing main exciter. If this exciter is relatively unstable changes may be necessary in the machine in order to provide stability under manual control. Moreover the manually operated main-exciter-field rheostat may be rather bulky and expensive particularly if a wide and low ranges of exciter voltage are to be controlled smoothly.

E. L. Harder (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The larger portion of this paper is devoted to a necessary statement of existing excitation and regulation principles, devices, and practices. Against this background, the newer developments are described, and their merits and field of application indicated. My comments pertain to the newer developments in excitation systems.

The authors have mentioned one type of rotating amplifier. Another type is the rototrol which is used to control the field of the main exciter. Its small control field energy requirements are supplied readily from an impedance or electronic type regulator.

The ultimate in simplicity of an excitation system is obtained with the main-exciter rototrol which eliminates the pilot or control machine altogether. A unit of this type is being built for a 60,000-kw turbo generator, a number of smaller units having been built in the past. The main exciter rototrol can be controlled by a static network or an electronic regulator.

The use of a lower excitation limit for a generator or frequency changer is recommended highly. This limit preferably should be variable with the load being carried, but in any event should be adequate to insure steady state stability for existing or increment loads. The provision of sufficient field current for stable operation is always paramount to variation of field current for voltage or wattless power control. The watt-balance element has been used for this purpose with indirect acting regulators. It balances slip-ring voltage against power through the unit, taking control away from the voltage regulator and increasing the field current if needed to insure the predetermined margin of stability.

J. B. McClure, S. I. Whittlesey, and M. E. Hartman: The authors wish to thank all those who have submitted discussions on this paper. While there seems to be general agreement on this subject, there are a few items which profitably can be emphasized.

Kinghorn and Porter have suggested that the electronic pilot exciter be included in the list of excitation systems. The electronic pilot exciter was considered when making up the list, but it was not included since the list was intended to be a summary of present day practice and the electronic pilot exciter has not been used widely by the industry. However, the experience with the electronic pilot exciter in a few installations has been quite satisfactory.

Three important characteristics are given which should be considered in selecting voltage regulators. In discussing accuracy it should be noted that regulators with no inherent dead band, in general, will have a higher accuracy because they are continuously acting. Regulators of this type are built with different degrees of accuracy which is reflected in their cost. Speed of response sometimes is confused with the continuously acting principle. Moderate speeds of response generally are considered adequate for most applications. One of the principle requirements of voltage regulators is that they begin to make corrections promptly. This requirement can be met by a regulator with a small dead band.

Harder indicates that there are several types of rotating amplifiers. The term rotating amplifier as used in the paper was intended to include all equipments of this type such as the amplidyne, rototrol, regulex.

Harder states that the simplest excitation system would make use of the main exciter rototrol. Main exciter amplidynes also have been built and now are being used in several smaller installations. However, such special exciters are necessarily larger. Also, the single special exciter requires higher control energy than if another amplifying stage is included. This in turn means greater duty on the regulator. In an excitation system including a regular exciter and a separate rotating amplifier, it is possible to remove the rotating amplifier from service without shutting down the exciter.

Valentine and Harder both indicate that excitation limits are worthwhile. It should be understood that the lower excitation limit protects the system by maintaining a minimum excitation for each machine, while the upper excitation limit protects the machine from excessive currents. Mr. Valentine mentions the use of an overcurrent limit on synchronous condensers but does not state whether this limit is provided to protect the machine against excessive armature or field currents.

Valentine also discusses briefly the modernization of old excitation systems. In many cases the exciter is relatively unstable under manual control. A small booster voltage, as mentioned by Kinghorn, will improve the stability characteristics of an exciter and increase the speed of response at low excitations.

Six Years' Experience With Factory-Built Unit-Type Substations

Discussion and authors' closure of paper 46-176 by E. F. Miller and R. B. Miller, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTONS, 1946, November section, pages 751–6.

Raymond J. Salsbury (Duquesne Light Company, Pittsburgh, Pa.): The adoption of a primary network is decided by economics. There is little doubt as to the desirability of the network from an engineering standpoint. However, there are certain details which could bear elaboration.

It would be desirable to present information as to the degree of regulation to be expected from the network application and as to what extent these expectations would be realized. Inverse reactance compensation is necessary to insure proper parallel operation of the network units. With a network partially loaded, the regulation problem is not difficult. However, it may be that when the network is loaded fully, bus regulation with the inverse characteristic will not be adequate.

Lightning arresters should be placed as close to the terminals of the transformer unit as is physically possible, regardless as to whether the entrance is overhead wire or cable. Open terminal bushings on the transformer unit obviously can be protected more easily with arresters adjacent to the bushing. The flashover mentioned in the paper must have been a result of some difficulty with the insulation of the bushing or the breakdown characteristics of the arrester. Bushing flashovers of the type suggested have been eliminated almost completely from many power systems.

Further discussion of some equipment details would be of interest. For instance, are the recording ammeters of the instantaneous or thermal type? What policy of loading is applied to the network units and what type of supervisory indication or remote control is provided, if any? Many of the maintenance features of the unit station could be treated more fully. The cables which supply the individual units will fail. How are the individual phase conductors identified when the line crew is required to repair a failure? What methods are used to insure proper phase rotation when a repaired cable is energized again on the system? Perhaps this problem is avoided through the use of single conductor cable in the supply.

The frequency of major inspection and overhaul is of prime importance. While the units are new, little trouble may be expected from them for several years. However, at some time transformer windings will have absorbed moisture and will require drying in a vacuum, the tap changer contacts will have worn and require replacement, the oil circuit breaker will have become obsolete and will require major rebuilding, and many of the other devices will require replacement. Furthermore, when something goes wrong inside the tank, the whole equipment will be out of service until repairs are made. It seems then that the following questions are pertinent: What provision is made for major equipment replacement or overhaul? How long will a unit give reliable service without any attention whatsoever? What reserve capacity is provided in the design of the system?

Finally, one cannot be too optimistic about the over-all economy of the primary network as compared with the radial scheme. Substantial distribution circuit investment is required to minimize regulation. This may be offset by substation economy. However, after these new unit stations have been in service 15 or 20 years, and there have been internal failures which require expensive replacement, many of the initial favorable factors may lose some of their value.

O. B. Falls, Jr. (General Electric Company, Schenectady, N. Y.): The authors have presented a very fine exposition of the use of and experience with unit type substation equipment and the advantages and economies of the primary network as contrasted to the radial type of system. This discussion will be confined chiefly to the latter item.

There are a number of reasons why the use of several relatively small substations is considerably better than the use of one large substation to serve a given area. Among these, as the authors have pointed out, are the problems of congestion around the large substations when a large number of 4-kv feeders must be taken away, and the necessity of supplying these large substations with a number of high capacity subtransmission lines.

Another of the major advantages also is mentioned by the authors. However, I would like to take this opportunity to further emphasize it. This pertains to the interrupting capacity of substation circuit breakers. Past experience has shown that one of the first items of equipment which caused a substation to become obsolete has been feeder circuit breakers. This was a result of an attempt to increase continually the capacity of these substations, which meant that, as the system grew, existing circuit breakers would reach not only their interrupting capacity but also their thermal capacity quite rapidly. The use of small equipments spotted at a number of locations throughout the load area and operated either radially or tied into a primary network means that the circuit breakers, from both an interrupting capacity and thermal capacity standpoint, will be adequate throughout the life of the substation as a whole.

Along this line, the question frequently is raised as to whether or not circuit breakers when used in a primary network will remain adequate as the network grows. Experience has shown that even in primary networks of quite high load densities and with high values of available short-circuit capacities on the subtransmission circuits, circuit breakers will remain adequate. Assuming a 4.16-kv primary network, the borderline where a 150-megavolt-ampere circuit breaker begins to become inadequate is at a load density where serious consideration should be given to installing a secondary network. Because primary networks usually are limited to residential, and mixed residential and commercial areas, it is unlikely that load densities will reach such values that any 150-megavolt-ampere circuit breaker would become inadequate based on present-day or expected future loads in such areas.

For example, if we assume an infinite network made up of 3,000-kva transformers and with 500-megavolt-ampere short-circuit capacity available at the high-voltage terminals of each transformer, the maximum short-circuit capacity to which any circuit breaker will be subjected on a 3-phase fault will be approximately 105,000 kva if the unit substations are spaced uniformly, 0.5 mile apart, and tied together with 3-phase overhead circuits of 0000 conductor. This means that a 150-megavolt-ampere circuit breaker with a reclosing time of 10 or 15 seconds on the first shot could be used with adequate margin. If the substations are spaced approximately 0.6 mile apart, a 150-megavolt-ampere circuit breaker could be used with zero reclosing time on the first reclosure. From this it can be seen that 3,000-kva transformers with 150-megavolt-ampere circuit breakers could be used for load densities in the order of 10,000 to 12,000 kva per square mile without becoming inadequate. This is based on assuming zero time for initial reclosure. If longer times are used on initial reclosure higher load density areas could be served.

The authors have pointed out that after careful consideration of the geographical layout of the area, the cost per kilovolt-ampere of the various sizes of substations, load density, and spacing of substations, a unit rated 3,000-kva transformer with forced air cooling to give it supplemental overload rating of 3,900 kva, was chosen as a size which would give the lowest over-all system cost. Such a conclusion lines up with the results which were obtained from an over-all generalized system analysis.[1] In this analysis, it was shown that when considering such factors as voltage conditions, continuity of service, size of load area and load density, subtransmission voltages, and size of substations, the general conclusion could be reached that in the majority of cases the 3,000-kva self-cooled rating of transformer when used in a primary network would give the lowest over-all system cost. It is gratifying to note that in this very practical application, identical results were obtained.

The authors have enumerated a number of advantages of the primary network as contrasted to other types of systems including the radial type of system. I would like to take this opportunity to stress a few of these advantages which I do not believe have been appreciated fully in the past. One advantage is that the use of 3-phase 4-kv lines connected in the primary network not only results in lower losses but also in better voltage conditions being obtained than could be obtained with single-phase feeder-voltage regulation on a radial system.

In connection with this, there is one further factor which the authors have not mentioned, but which will become increasingly important. That is, the use of tie circuits as obtained in a primary network will permit the use of larger motor sizes without obtaining objectionable light flicker. This becomes more important as the use of room coolers and air-conditioning equipment increases.

The second item which I would like to mention is that the authors have found that the operating and maintenance costs of a primary network actually have been less than for a comparable radial system. One of the outstanding reasons for this is that substations including circuit breakers can be maintained adequately without the work and complications of either installing a transfer bus or cutting over the loads to other substations.

In conclusion, I would recommend this paper quite highly to those who desire a primer on the advantages of a primary network, as well as the results which can be obtained, from a system planning viewpoint, by including and analyzing a wide variation of the various factors involved in system design before arriving at the size of substation to be used.

Reference

1. A Method of Approach to the Consideration of Power-Distribution-System Costs, O. B. Falls, Jr. AIEE Transactions, volume 63, 1944, pages 1078–84.

P. R. Pierson (Westinghouse Electric Corportation, East Pittsburgh, Pa.): The authors are to be complimented on the subject presented in this paper. It is timely because of the trend in this type of construction at the present time. It provides the answers to questions of suitability in an unquestionable manner. Together with the authors who had the courage to pioneer a program as outlined, the manufacturers also can derive satisfaction from the fact that these pioneering switchgear designs have given satisfactory service.

Three points of particular interest are that these installations are all air-circuit-breaker metal-clad switchgear, that the installations are of the simplest basic unit construction, and that the locations in the industrial centers of the Quad Cities on the banks of the Mississippi River are subject to as severe atmospheric conditions as usually will be encountered.

Drawout air-circuit-breaker metal-clad switchgear is adapted particularly well to unit substation or outdoor switchgear because, in addition to the safety features of metal-clad construction, the circuit breakers are individually mobile, thus requiring no auxiliary devices for operation within the unit. The circuit breaker parts are located in a position to provide for easy and quick inspection by the removal of a single piece barrier assembly. These operations and minor maintenance can be performed with the circuit breaker in front of its housing without the use of auxiliary handling devices other than a platform dolly.

The units discussed employ the simplest circuit construction. No provision is made for by-passing circuit breakers or transferring circuit to transfer busses. Thus a minimum of primary circuit subject to fault or maintenance is involved in each housing; however, maintenance has been performed satisfactorily without undue annoyance to customers. It is acknowledged that the network system contributes greatly to the continuity of service during these operations.

The six years of reasonably trouble-free operation of this type of equipment in the severe atmospheric conditions of the industrial sections along the river confirm the moisture prevention features of the switchgear design for most locations. The use of free ventilation through all compartments and the provision of heaters to maintain temperatures inside the housings above those outside during the severely changeable weather of spring and fall has proved adequate for the conditions.

The trend in unit substation switchgear is toward better appearing units, and standardization of circuits, equipment, and individual unit designs. With standard individual units, expansion, reduction or rearrangement can be made as load trends demand, with maximum salvage value of equipment.

E. F. Miller and R. B. Miller: In our opinion, good voltage regulation is one of the advantages of the primary network form of distribution. This is a result of the over-all design of a complete network. We have found that for our loadings, it is necessary to use 0000 conductor wire on the tie lines to sustain the network with one of the transformers out of service during a peak load condition. Conductors of 0000 wire on a standard 4-pin arm have an impedance of less than 0.7 ohm per mile. With tie lines two miles and less in length, the voltage drop from the station bus to the center of the feeders will be about two or two and one-half per cent, which is small in comparison to the drop in the secondaries and distribution transformers.

We set the compensator in each transformer tap changer control for a slight rise in voltage on its station bus as the load increases, with a maximum of about 121 secondary volts. We find that this gives excellent primary regulation throughout the day and night and makes the matter of voltage at the customer's entrance a function of the distribution transformer and secondary drops.

These results are based on maintaining balanced loads on the individual phases of each circuit and also on the basis of using four circuits per substation with two or perhaps three of them as tie lines. The primary drops referred to are based on about 2,800-kva normal peak load on a 3,000 3,900-kva substation and with power factor of 80 per cent or better.

A network of this general design is capable of maintaining good operating voltages throughout on peak load with any of its transformers out; also, any tie line will function satisfactorily with one end open. This permits maintenance to be done on a circuit breaker or transformer any time without inconvenience to the customer, and during regular working hours. With uniform design of substations one spare circuit breaker and other important pieces of equipment can be provided for use anywhere on the system. A circuit breaker unit needing major work done on it can be replaced quickly and taken to a service building for major testing or repair. Minor jobs can be done at the substation.

The use of air circuit breakers instead of oil circuit breakers minimizes the amount of maintenance of such units. Transformers have automatic pressure gas seal above the oil and cuts oil contamination to a minimum. This reduces maintenance and possibility of trouble besides extending the expected life indefinitely. Weekly routine inspection trips catch all the minor trouble and prevent major trouble from developing.

All substations have instantaneous recording ammeters, wattmeters, and varmeters besides a watt-hour meter. These instruments record quantities for the transformer as a whole. Test jacks are provided on each circuit for connecting portable instruments for spot checks or major surveys. There is no supervisory control or indication.

Relaying is provided to trip the transformer's secondary circuit breaker in case of a fault on the primary circuit or in the transformer winding. These relays provide automatic reclosing after the primary circuit has become energized, and the magnitude and phase angle of the voltage is such that it will feed energy into the network. Relays also are provided to trip all circuit breakers in case of a substation bus fault. Each feeder has overcurrent relays on it. The transformer has a winding temperature relay set to trip the transformer circuit breaker if a safe temperature is exceeded.

Single conductor cables are used on all 4-kv and 13.2-kv lines into each substation. On 4-kv circuits the phases are marked at the switchgear and at the pole for ease of phasing out in case of cable trouble. Distribution type lightning arresters are provided at each dead end when there is a cable terminal.

Some Aspects of Polyphase Motor Design—The Design and Properties of the Magnetic Circuit

Discussion of paper 46-166 by T. C. Lloyd and H. B. Stone, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, December section, pages 812–18.

P. H. Trickey (Diehl Manufacturing Company, Somerville, N. J.): T. C. Lloyd and H. B. Stone are to be complimented for a very clear analysis of an old and difficult design problem, and particularly for their ability in eliminating the confusion and complication of certain awkward and comparatively unimportant factors.

Their formula for tooth width checks very closely with my own experience. The equation for inside diameter checks with the curve shown in Figure 3 of my discussion on Lloyd's 1944 paper, for a stator diameter of 4.6 inches. However, perhaps Lloyd's formula should be modified for subfractional sizes because it would give zero rotor diameter for a 0.647 diameter stator.

Very often it is not possible to use a rotor diameter exactly as desired because of the need to use previous die diameters. In this case it may be better to choose the "depth below slot" from a proportional division of the distance from inside to outside diameter. Such a formula is

$$DBS = \frac{0.86}{0.315P+1.03}\left(\frac{OD-D}{2}\right)$$

A further extension of the principles of this paper might be to choose the rotor diameters so that the 4-pole diameter of one frame size would be the 2-pole diameter of the next larger frame, and the 6-pole diameter of the next smaller frame.

Characteristics of Resistance Welding Transformers

Discussion of paper 46-156 by F. E. Murray and R. C. Jones, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 726–8.

F. L. Brandt (nonmember; Thomson Electric Welder Company, Lynn, Mass.): I am pleased to see how close the authors' design constants agree with my own. This is true because our designs are governed by accumulated experience of these transformers in service.

I would like to amplify the data given on core design. Equation 3 assumes a stacking factor of 0.95. This stacking factor expresses a sort of efficiency in the way that the core is cut and assembled. I would like to point out that the dies or shears used to cut the laminations must be maintained religiously so that burring of the edges of the laminations is at a minimum. The core assembly personnel must stack these laminations so that there is a minimum of air spaces. The laminations should be cut so that the flux path is in the same direction that the steel has been rolled. These points apply to all transformers and not just resistance welding transformers.

There are silicon steel cores available that can be worked up to 115,000 to 125,000 flux lines per square inch. A large part of this increased density is enabled by the method of assembling the core laminations which prompted the discussion in the previous paragraph. Such cores are costly in themselves but the saving in primary and secondary copper resulting from reduced length of the mean turn actually reduce the total cost of the transformer in many cases. Not only the physical size of the transformer is reduced but so are the losses and impedance. Higher flux density cores are a definite trend for the future.

I should like to point out that the high voltage windings should be insulated with a suitable varnish or impregnant to resist moisture, oils, and acids that are prevalent in manufacturing plants. The transformer is part of a machine in which lubricants often are used freely. The water-cooled low-voltage windings are apt to sweat like a pitcher of ice water on a hot summer day.

There is a growing trend to make the low voltage windings of rolled copper, particularly in the form of heavy wall tubing. This further reduces the size of a given transformer.

The leakage reactance and losses are calculated for the parallel connection of the 100-kva transformer.

Moisture Equilibrium Between Gas Space and Fibrous Materials in Enclosed Electric Equipment

Discussion and author's closure of paper 46-160 by John D. Piper, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, December section, pages 791–7.

F. J. Vogel (Illinois Institute of Technology, Chicago, Ill.): This paper is of great interest to the designer and manufacturer of apparatus inasmuch as it points out one of the problems that has to be considered by the designer which generally is not appreciated by other engineers.

The author says that he does not know of transformer failures from condensation. There have been several. This condensate has not always been water only but in some cases a mixture of acid and water. In my opinion, it has caused both terminal structure failures and one or two insulation failures in transformers.

That condensation has been given at least a little consideration is shown by several schemes, some patented, for providing extremely free breathing to prevent condensation. Free breathing also accelerates oil deterioration caused by oxidation.

The presence of condensation has been used to determine dryness. Units have been heated by circulating current in the windings and then examined to note any condensate on the underside of the cover. Absence of moisture on the cover is taken as an indication of dryness. However, these tests are not quantitative nor are they necessarily uniform in their results.

Although new oil has an extremely small percentage of water solubility, this may be a large quantity of water if condensed from a large transformer.

This solubility decreases with lower temperature and undoubtedly the oil volume acts within limits as a water reservoir. The water then is distributed between the insulating materials, oil, and the gas space of sealed units. Where the temperature does not range over extreme limits, there is probably little trouble. From experience, and not as a result of test, it might be concluded that the dielectric strength of insulations is not affected greatly by at least as much water as would be picked up by contact with the air. However, if a transformer were heated up to about 50 degrees centigrade or more in a moist atmosphere, considerable moisture might go into solution in the oil. If the unit then were placed outdoors at minus ten degrees centigrade the oil solubility could be reduced to the point where there would be freely divided water in the body of the oil. Then there is likely to be free water on the surface of the insulation, and cases where it could have been the cause of trouble are known. Piper has given consideration to these problems, shown when he states that the equilibrium for transformers is more complicated than the tapes he tested. Although I believe safe procedures are known and prescribed by the manufacturer's instructions, I am not aware of exact data that are available.

These phenomena are also of interest in the drying process. Drying is obtained by heating or the application of vacuum. The time and the heat required to vaporize the moisture are elements of this problem and consideration has been given to studies of this subject.

C. T. Hatcher (Consolidated Edison Company of New York, Inc., New York, N. Y.): The author has presented a comprehensive paper on the question of determining moisture in materials. There is one question which requires an explanation concerning the partial pressure of unsaturated water vapor. It is stated that for any vapor pressure long continued exposure of any fibrous material will cause it to absorb or give off mositure until its moisture content is in equilibrium with the vapor in the air. From example 2 it appears that the writer assumes the pressure exerted by unsaturated water pressure at any temperature to be the same as the pressure exerted by the same amount of vapor per unit volume at saturation. Actually, it would be somewhat higher because of the fact that it is superheated or raised to a temperature above that of saturated water vapor of the specified density. This will make no difference if the curves showing moisture content of fibrous material at different temperatures are based upon the same conception, but it is not clear whether or not this is the case. If it is not, an error of about eight per cent in the moisture content of the fiber at equilibrium would result.

Abe Tilles (Pacific Gas and Electric Company, San Francisco, Calif.): As is well known, cable engineers and others interested in maintaining the integrity of high voltage insulation always have a great and serious concern regarding the whereabouts of the last traces of moisture. Accordingly, great interest is attached to the method, so ably developed here, of establishing moisture concentrations in relatively inaccessible dielectric materials from measurements on readily reached gases in contact with them.

The author states that equilibrium times are quite long. Can the time be given more specifically of approximately how long it might take to reach equilibrium? If this time is long compared to that of a marked daily temperature cycle, what temperature should be considered as controlling—the maximum or average temperature of the insulation surface, of the insulation volume, or what?

Following the reasoning of the paper further, would it be feasible to dry the insulation slowly but surely by leaving some highly hygroscopic material in contact with the gas space? Has this been attempted?

Would it not be expected that the presence of various true gases, for example, nitrogen with or without oxygen, at various gas pressures over a wide range, for example, two millimeters of mercury up to 15 atmospheres, would leave the moisture equilibrium concentrations unaltered?

The authors state that presence of a thick impregnant does not alter moisture equilibrium concentrations of gas space and fiber. When, however, a volatile and hygroscopic vapor—as perhaps of transformer oil—is present, would they not expect the equilibrium moisture content of the gas and vapor space to be subject to marked alteration on that account?

L. Meyerhoff (General Cable Corporation, Bayonne, N. J.): J. D. Piper very cleverly has drawn together the scattered data of equilibrium moisture conditions and put these into a very convenient and usable form. He has shown some practical confirmation and has indicated various applications for which these data will be valuable.

It is of interest to compare with his curves, equilibrium data we have taken of unimpregnated kraft paper in the range of 15 to 70 per cent relative humidity.

In our tests four sheets of cable paper of different densities, each weighing approximately 50 grams, were hung up in the laboratory and thus subjected to the laboratory atmosphere. The initial moisture content was determined by drying other similar samples at high temperature to constant weight. The samples were weighed at frequent intervals during the 14 months of the test (January 1932 to March 1933) and at the same time temperature and relative humidity were observed. It will be noted that the paper did not necessarily attain equilibrium at any one atmospheric condition. However, some readings were with rising relative humidity and some with dropping relative humidity. Thus the average moisture content observed for a given atmospheric condition is probably not far from the equilibrium value.

Table I

Per Cent Relative Humidity	Per Cent Moisture Content	
	GCC* Data	Data From Figure 4
15	4.2	4.1
20	4.6	4.9
30	5.5	6.3
40	6.3	7.6
50	7.2	9.3
60	8.0	10.1
70	8.9	11.2

* General Cable Corporation.

The degree of accuracy of the method does not permit determination of the effect of temperature and humidity independently of each other. However, it is not too far out of line to consider the data as fairly accurate for the average temperature in the tests, 26 degrees centigrade, with a range of 22 to 32 degrees centigrade No significant differences were found among the various samples because of density, including superdense paper.

The average results are given in Table I of this discussion in comparison with values taken from Piper's curves for 26 degrees centigrade.

The comparison for low relative humidities is extremely good and is not far out of line for the higher humidities.

With saturated paper the rate of moisture absorption may in some cases predominate in practical importance over the final equilibrium condition. Here the question of degree of saturation and, in an extremely important degree, the type of saturant become controlling and the rate of absorption is only a small fraction as much with certain saturants as for the unimpregnated paper. The surrounding humidity greatly affects the rate of absorption as well as the equilibrium point.

A rapid device for determining the effect of exposure of saturated paper to moisture is by tape-by-tape power factor measurements. These are much more effective when made at 80 degrees centigrade than when made at 60 degrees centigrade. Figure 1 of this discussion shows a comparison which we have obtained between power factor at 60 and 80 degrees centigrade of saturated tapes wherein the major variable is the amount of moisture present. The data are for an experimental cable which was taped with presaturated tapes in a room of relatively high humidity. Some of the tapes had been exposed to the humid air for considerable lengths of time during interruptions of operation. It will be noted that where the power factor is low (that is, very low moisture content) the 60- and 80-degree-centigrade power factors are almost identical, whereas for high power factor values (that is, relatively high moisture content) the 80-degree-centigrade power factor goes up much more rapidly than does the power factor at 60 degrees centigrade.

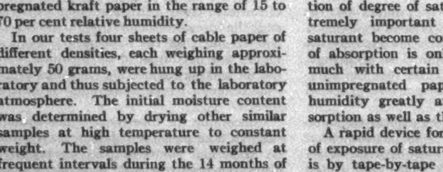


Figure 1. Comparison of 60- and 80 - degree - centigrade tape-by-tape power factor with moisture as the major variable

The dotted line shows one-to-one variation for comparison

The great protection (greatly slowing of absorption rate) obtained from the metallized paper shielding tape, as noted by Piper, is pertinent to this whole matter and has contributed in a great degree to the low moisture content shown by him for cable as completed and shipped.

E. W. Greenfield (Anaconda Wire and Cable Company, Hastings-on-Hudson, N. Y.): J. D. Piper's interesting co-ordination of previous work on adsorption isotherms for cotton, paper, and wood gives a unique dew point chart for cellulosic insulating materials.

Figure 4 of the paper permits determination of equilibrium moisture content of cotton or wood pulp paper for any given temperature and vapor pressure of its surroundings. Since the curve for water also is given, the chart can be used to find dew point temperature corresponding to various operating conditions.

The value of such an equilibrium chart becomes apparent from consideration of the examples given by Piper. Other problems involving the equilibrium moisture content for paper, varnished cloth, fabric tapes, and so forth stored at various relative humidities are of equal interest.

Construction of the vapor pressure curves for wood pulp paper based on the data of Pidgeon and Maas (reference 8 of the paper) may be open to some question. It is true that the cellulose micelle in any state maintains its unit chain individuality since cellulose from many differing sources shows identical X-ray diagrams. (These diagrams are also identical for cellulose both in the very dry and very moist states indicating the occluded water is not bound within the unit cells but probably exists entirely as surface masses.) However, the lignon and resin impurities for the spruce are much greater than for normally-manufactured wood-pulp paper. Also, the oxycellulose content differs appreciably and oxycellulose has a greater affinity for water than alphacellulose. Piper's extrapolation gives him a ratio of 1 to 1.7 for the relationship between the water concentrations that are required to produce the same vapor pressures in cotton and in kraft paper. We feel that this ratio is too high and is occasioned by the assumption of the identity of behavior between spruce wood and wood pulp paper.

We are in agreement with Piper in that the data on unimpregnated cellulose is aplicable to oil-impregnated cellulose. Although the presence of the saturant slows up the rate of diffusion, it would appear that the moisture gradually displaces the saturant and becomes adsorbed itself to the cellulose as surface masses. Tests on unimpregnated and impregnated kraft condenser tissue carried out at Bell Laboratories during 1939 indicated definitely that the same final moisture equilibriums are reached.

In his discussion on vapor barriers, Piper states that either metallized paper or metal tapes placed on the outside of the cable will form a positive barrier to entrance of water vapor. Has consideration been given to the long spiral exposed cellulose edge in the case of the metallized paper or the greater exposed cellulose surfaces of the intercalating tape used with all metal shielding tapes?

G. B. Shanklin (General Electric Company, Schenectady, N. Y.): The interesting information about the effects of moisture in high pressure pipe cable systems, as outlined in J. D. Piper's paper, is appreciated. This experience is further confirmation of the old basic principle of high voltage cable practice and applies equally well to any type of high voltage system, either lead sheathed or in steel pipe. In all such systems it is essential to remove all possible traces of impurities, such as oxygen and moisture, from the cable initially and to safeguard against trapping such impurities in the system during installation and service.

In this particular case the moisture was entrapped by exposure before the temporary lead sheath was applied. There was plenty of time for this moisture to become firmly embedded in the cable cross section, and the cold vacuum treatment applied in the field after the cable was installed was not sufficient to remove this deeply embedded impurity. In this respect the limitations of any field treatment it is practical to apply must be recognized, and penetration of oxygen and moisture into the mass of insulation must be avoided. In the case of pipe cable systems, where the temporary lead sheath is stripped off as the cable is pulled into the pipe in the field, it is important to apply vacuum treatment as promptly as possible after the pulling and before the impurities caused by exposure have a chance to penetrate the mass of insulation. It is equally important to make sure that there are no similar impurities in the cable before the sheath is stripped. This is a problem we always will have as long as paper-insulated cable is used.

W. E. Mougey (Bell Telephone Laboratories, Murray Hill, N. J.): We are not sufficiently familiar with recent developments in power cable to say whether the air pressure cable is finding enough acceptance to warrant the attention given it by J. D. Piper. It is our impression, however, that there has been considerable interest in this development. We assume also that there is rather wide interest in oil filled transformers of the types considered.

While Piper does a lot of extrapolating and interpolating, we can find nothing basically wrong with his methods. For the most part his extrapolations are based on linear or nearly linear curves. However, we are somewhat skeptical of one result of his extrapolations, namely the conclusion that kraft paper takes up 1.7 times as much water as cotton. While we agree that kraft paper is more hydroscopic than cotton, we are surprised that the factor is as high as 1.7. It would be a simple matter for the author to make moisture absorption measurements on kraft paper and cotton under conditions to bolster his conclusions.

Table II

Per Cent Relative Humidity	Per Cent Moisture Content		
	GCC* Data	ICT** Data	Figure 4
15	4.2	4.0	4.1
20	4.6	4.8	4.9
30	5.5	5.7	6.3
40	6.3	6.7	7.6
50	7.2	7.7	9.3
60	8.0	9.0	10.1
70	8.9	10.6	11.2

* General Cable Corporation.

**International Critical Tables.

A good job has been done by Piper in rationalizing the moisture absorption data of a number of investigators. This is probably one of the most important features of the paper.

In the discussion of the moisture takeup of impregnated versus unimpregnated fibrous materials, we suggest that reference be made to an article by Ingmanson and Vacca.[1] Data are given there showing that the equilibrium moisture content of jute is not changed by impregnation.

REFERENCE

1. MOISTURE ABSORPTION OF JUTE, J. H. Ingmanson, G. N. Vacca. *Industrial and Engineering Chemistry* (Easton, Pa.), volume 26, 1934, page 1274.

John D. Piper: The writer appreciates the interest in the paper expressed by the several commentators and is gratified by the opinions of E. W. Greenfield, L. Meyerhoff, Abe Tilles, and F. J. Vogel that the information compiled may be put into practical use.

Greenfield and Mougey express the opinion that the ratio of 1.7 to 1 is somewhat high for the relationship between water concentrations in kraft paper and cotton required to produce the same vapor pressure.

Meyerhoff presents experimental evidence which substantiates the conclusion that for the higher humidities the ratio is high, but which confirms the values from Figure 4 for the lower humidities. Meyerhoff's data are substantiated qualitatively by data taken from "International Critical Tables" as shown in Table II of this discussion.[1]

Divergence from the International Critical Tables data is, however, less than that from the General Cable Corporation data. In preparing the equilibrium chart the writer was forced to choose between data for kraft paper taken at room temperature and data for spruce wood taken over a range of temperatures. The latter was chosen in spite of the differences between kraft paper and spruce wood which ably have been pointed out by Greenfield. Except at high humidities the agreement with the meagre data for kraft paper seems remarkable. For precise work an obvious need exists for further data on kraft paper taken over an extensive range of both temperature and moisture concentration.

In response to Hatcher's comment, it should be stated that the curves as given show the equilibrium conditions against water vapor of the vapor pressures shown. Hatcher rightfully points out that in the illustrations the effect of the superheating was neglected. Assuming the gas laws to hold, the equilibrium vapor pressure for examples 2 and 3 as corrected for the superheating is 6.3 millimeters instead of the uncorrected value of 5.7 millimeters of mercury.

Regarding Tilles' opinion that true gases such as nitrogen should not affect the equilibrium conditions but that volatile hygroscopic materials such as transformer oil might cause marked alterations, the writer partially agrees. Experimental evidence shows that nitrogen pressures up to 200 pounds per square inch do not affect the equilibrium conditions beyond experimental error. Vapors of hygroscopic materials probably would if present in sufficient concentration. The odor of the gas over transformations is usually acrid indicating the presence of acid vapors. Analysis of condensate from particularly moist transformers showed, however, the acid to be present in proportion less than 0.5 mole per cent. These are unpublished data of the writer's colleague, T. A. McConnell. Inasmuch as the acid is less volatile than water the concentration of acid in the vapor phase was probably even less. The concentration of oil vapor also must have been low because 98 per cent by weight of the condensate was water. From Raoult's law it is evident that the equilibrium conditions cannot be altered seriously unless the concentrations of hygroscopic materials are much higher than were found in the case cited. In connection with this and the previous comment it should be pointed out that the information given in Figure 4 is not intended for precision work but for making engineering estimates over a wide range of temperatures and concentrations.

Concerning the time required for equilibrium to take place, few data are available. As Meyerhoff points out, for impregnated fibers the rate is greatly dependent upon the nature of the impregnant. For transformers and the cable wrapper discussed, sensible equilibrium seems to take place in a matter of days. Concerning the proper temperature to use when the period of a temperature cycle is less than that necessary for complete equilibrium to take place, it has been the practice in working with transformers to use the average temperature of the transformer oil during the working day.

Use of a highly hygroscopic material in contact with the gas or oil in transformers has been proposed in the form of several proprietary devices. Too many of these proposals neglect the fact that the equilibrium vapor pressures of these desiccants increase rapidly as moisture is absorbed and that the quantity of desiccant required to dry a large transformer is considerable. Nevertheless this approach has merit provided a degree of dryness approaching that in high voltage cable is not required. A compressor-condenser system is being used by the company with which the writer is connected on the gas system of 19 generator-transformers in which particularly high dew points were found. While in operation this

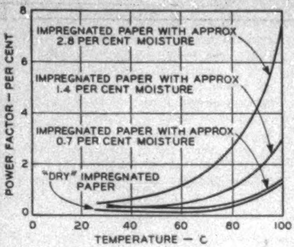


Figure 2. Effect on power factor of moisture in impregnated cable paper

From minutes of the Association of Edison Illuminating Companies for 1929, "Study of Mechanism of Cable Deterioration" by C. F. Hirshfeld, A. A. Meyer, and L. H. Connell

system prevents condensation except in coldest weather and is slowly reducing the moisture concentration in these transformers.

The writer is grateful to Greenfield and Mougey for calling attention to the prior work of Ingmanson and Vacca of the Bell Telephone Laboratories in showing the equilibrium moisture content of a fibrous material to be unaffected by the presence of an impregnant.

Meyerhoff's comment concerning the use of power factor measurements as a criterion of moisture content of impregnated insulation makes it advisable to call attention to the data of D. E. F. Thomas that is reproduced in Figure 2 of this closure. These data showed that power factor measurements, even at 80 degrees centigrade, are not a reliable criterion of the moisture content of impregnated paper when the moisture content is less than approximately 0.7 per cent based on the weight of the dry paper.

There is no disagreement with Shanklin's contention concerning the necessity for protecting cable insulation from contamination by moisture and other impurities. The inference that the moisture entrapped by the cotton protective tape became embedded in the cross section of the insulation is, however, erroneous. The average moisture concentration in the insulation of the cable in question was as low as that of other well-made high-voltage cable at the time it was installed. Moreover, experimental data show that the moisture from the gas space has not migrated into the insulation, a fact which also answers the question, raised by Greenfield, concerning whether the vapor might enter the insulation through the long spiral exposed cellulose edge of the metalized paper tape.

Shanklin is quite right in his contention that the field drying to which the cable was subjected could not remove water—had any entered the insulation. Figure 4 of the paper shows that with the cable at a temperature of 40 degrees centigrade and evacuated to 2 millimeters of mercury the equilibrium water content in kraft paper would be approximately 1.5 per cent, assuming of course that the residual pressure was entirely water vapor. The sole function of the treatment was to reduce the water concentration of the protective tapes to the extent that the dew point of the gas never could be reached at local cool spots in the line.

As stated in the paper, data concerning the cable line were released only because they show that the same equilibrium holds for impregnated fibers as for unimpregnated. The writer had not intended that the data be used to indicate either the merits or the demerits of a particular type of cable. Inasmuch as the question has been raised, however, it seems only fair to state that the results of tests and the outstanding performance of the cable line to date have upheld amply the faith that the proponents of this particular type of cable had in it.

REFERENCE

1. INTERNATIONAL CRITICAL TABLES, first edition. McGraw-Hill Book Company, Inc., New York, N. Y., volume 2, page 322, Figure 1.

Temperature Limits for Short-Time Overloads for Oil-Insulated Neutral Grounding Reactors and Transformers—II

Discussion and author's closure of paper 46-165 by V. M. Montsinger and J. E. Clem, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 966–73.

M. S. Oldacre (Commonwealth Edison Company, Chicago, Ill.): We are given new data in this paper for use in calculating short-time overloads of oil-insulated equipment that can be especially valuable in determining the overload possibilities under different conditions of load and time than the original design basis.

The primary purpose of the work described in the paper seems to be to provide background data for the proposed new standard for neutral grounding devices, that just has been submitted to the Standards committee for approval.

After reading the proposed standard, it seemed to be a logical document subject to only minor questioning, but then after subsequent reading of the paper by Montsinger and Clem it did not seem that there was full co-ordination of the data in the standard with that in the Montsinger and Clem paper.

It is suggested that careful study of the paper is required before full consideration can be given to the proposed standard.

This paper brings out another very valuable point and at the same time initiates a new practice in giving the temperature limits for loading of oil-insulated apparatus. This statement refers to the different temperature rises that are proposed for different starting temperatures at the time of the overloads. In effect a combination is made of the previous methods of rating apparatus and the correction in loading for different ambient temperatures. However, because it is new and different, it seems very essential that the ambient temperatures be set forth clearly when the temperature rises are stated.

One final point or question. Will Montsinger tell us if the data in the paper can be used for transformer windings of different sizes than mentioned? If it is generally applicable, it will be useful in many conditions that exist from day to day.

V. M. Montsinger: The only way in which I believe there is lacking full co-ordination of the data in the paper and in the standard is that the paper gives the amount of heat lost to the oil during a 1-minute fault whereas the standard does not. Because this amount is not very great, it is reserved as a factor of safety in the standard; that is, when calculating the temperature rise, only the heat stored in the copper and associated insulation is considered.

With reference to the last question, the data are applicable to windings of different sizes provided the construction is not entirely different; that is, it might not apply to a small high voltage design having an entirely different ratio of copper to associated (turn) insulation. It is rather doubtful if the data are applicable to the overloading of transformers in general for the reason that the time considered in the paper is too short (in the order of one minute) for the oil temperature to change as it does under usual overloading conditions. For short times in the order of one minute, the data apply.

An Electronic Drive for Windup Reels

Discussion of paper 46-125 by K. P. Puchlowski, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946 and published in AIEE TRANSACTIONS, 1946, August–September section, pages 585–91.

W. D. Cockrell (General Electric Company, Schenectady, N. Y.): This paper is greatly enhanced by the author's starting out with the simple formulas on which the drive is based. In fact, I would like to have seen also included the simple electrical formulas for motor speed and torque so that K. P. Puchlowski's circuit (which sets the speed to be held by setting the armature voltage) could be traced back quickly to the fundamental laws. By operating at the midpoint of the linear field range, a buildup of two to one should be obtained easily with good speed control. However, Puchlowski must have assumed fairly constant tension since a large change in tension would demand operation over an excessive field range.

It is assumed that this drive is not required to accelerate rapidly high inertia loads since field weakening produces a loss of torque when a higher speed is called for and thus the acceleration is diminished rather than assisted.

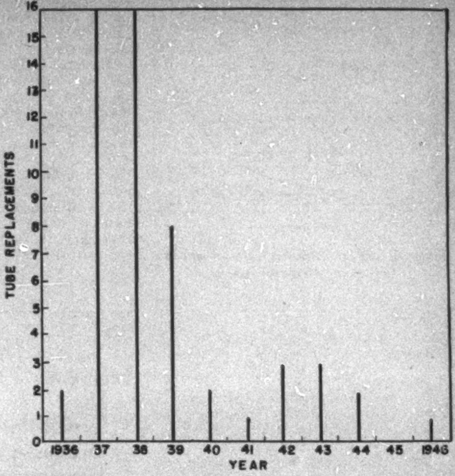
Although no attempt was made in the present drive to control the tension also, it might have been possible to do this by having a movable idler roll whose weight would maintain the tension and whose movement might be used to influence the *IR* drop compensation.

In the discussion of the basic control system, the statement is made that "the most accurate and sensitive regulator is unable to restore the deviated quantity exactly to its original value." This is true for a proportional action controller such as described, but a proportional plus reset action controller should be able to do this.

In the electric circuit itself, I am unhappy to see so many separate sources of direct voltage required. The circuit shows six batteries and two auxiliary rectifiers in addition to the rectifiers supplying armature and field. Not only do these indicate a high first cost, they also are a maintenance nuisance. Also, the circuit seems to contain a number of subcircuits which are strung together through what seem to be rather high impedances. This means that when the panel becomes slightly dirty and moist, as is almost unavoidable in most industrial plants, sneak circuits may appear which will be difficult to trace and which will cause erratic operation.

In conclusion, may I compliment Puchlowski highly on his use of the new standard graphical symbols in his wiring diagram. By standardizing on these symbols, we will produce diagrams that will be more understandable to both the electronic engineer and the industrial electrical engineer.

Figure 1. Chronological record showing sealed ignitron tube replacement frequency at the Fostoria, Ohio, exciter



The Development of Modern Excitation Systems for Synchronous Condensers and Generators

Discussion and authors' closure of paper 46-154 by F. M. Porter and J. H. Kinghorn, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1020–31.

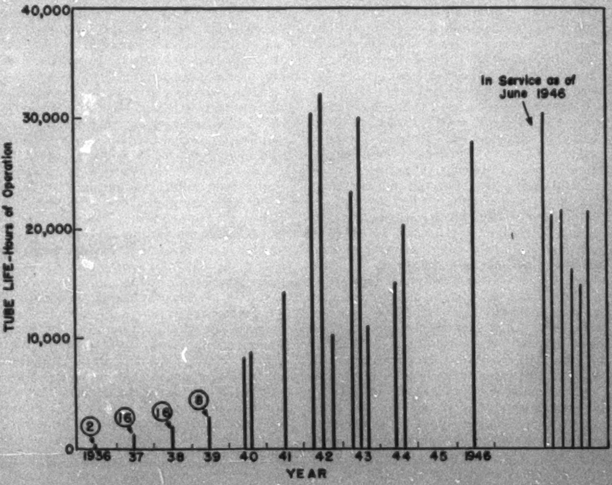
H. C. Steiner (General Electric Company, Schenectady, N. Y.): In commenting on the very excellent paper that F. M. Porter and J. H. Kinghorn have presented, I would like to discuss briefly one question that the authors have raised. It is the question of whether "the heavy expense of making tube replacements" in the electronic tube exciters can be reduced to a satisfactory level. Tube replacement expense depends, of course, on tube cost as well as tube life. In a technical discussion, however, we are limited properly to the technical and not the commercial phases of the matter. Therefore it is the purpose of this discussion to consider only the question of tube life.

In order to determine whether the authors' objective can be met, it is necessary, for the designer at least, to examine the individual rather than the average tube life. Figures 1, 2, 3, and 4 of this discussion have been prepared on this basis to show the tube life and tube replacement history of the Fostoria and Torrey exciters. All tubes that have been removed from the exciters are shown regardless of whether the cause of failure was the result of the tube design and manufacture or application. Fostoria represents the initial installation of this size and type of equipment while Torrey represents an almost duplicate unit in which the experience with the first unit was translated into the design of the second.

In the case of the Fostoria exciter the initial high tube replacement rate (Figure 1 of this discussion) and the low tube life (Figure 2 of this discussion) show clearly the result of the developmental work that was required as a result of field experience. In the development of those tubes in which a life of 1,000 hours is satisfactory, it is obvious that life tests can be run that in a relatively short time will allow the tube designer to make changes and to obtain satisfactory performance. As the life of the tube increases and particularly as the size increases, it becomes necessary to make actual installations in order to obtain the same experience. Forced life tests are useful only when the particular cause of failure is known and the test can be run for a par-

Figure 2. Ignitron power tube life at the Fostoria exciter

Individual tube life is shown except for the years 1936–39 where the average for the number of tubes indicated is given



ticular objective. As the life of a tube approaches three to five years, an equally long period will be required to show the results of any design changes. In the case of Fostoria, the individual life of the particular tubes indicates that a much higher average can be attained when the causes of failure in some of the shorter life tubes are removed. The individual lives of the tubes now in operation as shown by the bars at the right hand side of the chart bear out this conclusion.

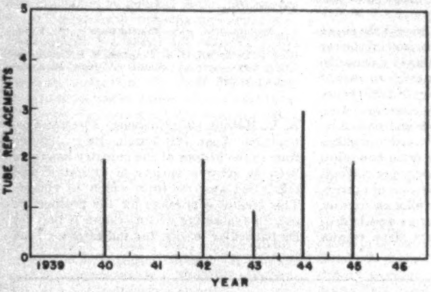


Figure 3. Chronological record showing sealed ignitron tube replacement frequency at the Torrey, Canton, Ohio, exciter

In the case of the Torrey exciter (Figures 3 and 4 of this discussion), the initial problems have been eliminated. The tube life is longer and the replacement rate lower. In the six and a half years of operation there have been 10 tube replacements or an average of $1^1/_2$ tubes per year. If we eliminate the two tubes (1942 and 1944) giving less than one year's service as manufacturing defects, the replacement rate is $1^1/_3$ tubes per year. This is permissible because the present tube warranty insures three years of service, or for an installation of six tubes replacement of not over two tubes per year.

The excellent life indicated for the tubes in service at New Carlisle also shows that extremely long life can be obtained under particular conditions of service. The tubes in this and in the Fostoria and Torrey exciters are of the same type and rating, and it is probable that the longer life is a result of the lower exciter kilowatts and higher voltage.

On the basis of the tube life obtained in the years 1942 and 1946, it is reasonable to expect an average life of four to five years. We would like to hear the authors' opinions as to whether this constitutes a satisfactory life or, if not, what would be satisfactory.

In regard to the general question as to whether a direct-connected exciter is preferable to a rectifier, the preference would seem to rest primarily on the particular conditions of the installation. Rectifiers for industrial, mining, and electrochemical service usually can be justified on the basis of efficiency, but in this case efficiency is of less importance than performance. Rectifier designs have been improved and standardized during the past decade, and as a result they are considerably simpler and less costly than the original exciters. It is possible that direct-connected exciters may prove more satisfactory on slow speed machines, and rectifiers more satisfactory on high speed machines where there are difficulties with mechanical commutators.

In conclusion may I express our appreciation of the co-operation that we received from the engineers of the American Gas and Electric Company and the engineers and operators of the Ohio Power Company in the development of the ignitron tubes and the electronic exciter. It look a great deal of faith to believe that the tube performance as demonstrated by the initial years at Fostoria could be made into anything approaching the reliability required, and in this period we received only co-operation and encouragement.

C. E. Valentine (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The report of operating experience as given by F. M. Porter and J. H. Kinghorn covering a period of several years provides many items of interest for both design and operating engineers. The variety of generating, excitation, and regulating systems which are described affords comparative information not available elsewhere.

For many years one of the main questions in the minds of designers and operators, with reference to electronic regulating devices, has been the question of tube life, and maintenance or replacement costs. This has been difficult to predict because service operating conditions are variable, and factory-made life tests can be completed at reasonable expense only on the basis of a small number of circuit conditions and operating conditions. The situation is in some respects similar to the testing of lamps in the lamp industry. Predetermined and accurately controlled overvoltage testing for a short time gives indication of life expectancy but a number of variables enter into actual lamp life, such as changes in voltage, temperature, or degree of vibration, and these make the actual life vary considerably from the life expectancy. Furthermore, improvement in materials, manufacturing methods, and testing take place year by year. In electronic devices these improvements have been accelerated greatly in recent years, and this points to improved performance in the immediate future.

The authors have stated that the initial cost of electronic exciter equipment and the heavy expense of making tube replacements offer objections to the use of this otherwise most desirable equipment; also that the performance of a rotary amplifier control system with direct-connected exciter has been essentially as good as that obtained from electronic main or electronic pilot excitation. It thus appears that initial tube cost and tube replacement cost has indicated the desirability of the rotary amplifier system. It is noted however that this rotary amplifier system employs an electronic voltage regulator and this is considered an ideal arrangment for a synchronous condenser, with due regard to factors of reliability, performance, economy, and least maintenance.

In their discussion of electronic regulators the authors state that an ideal regulator would have no moving parts and would require no periodic adjustment or renewal of parts. This seems to boil down to the inherent reliability of regulator component parts. It becomes a matter of the performance and life characteristics of transformers, resistors, reactors, and rectifiers compared with replacable components such as electronic tubes. The absence of moving parts does not make necessarily for an ideal regulator. Much depends on the simplicity and durability of moving parts, if used. As an example, spring mounting of such parts avoids wear and consequent readjustment or replacement. There are definite advantages in the use of simple regulator and machine field circuits. In case trouble does develop less skill is required and less time is lost shooting trouble where relatively simple devices and circuits are involved.

With regard to the electronic pilot and electronic main exciter, the authors have tabulated tube failures and tube life. With regard to the electronic regulators, however, it is stated that the tubes used in these regulators have given long life and their replacement cost is low. It would be interesting to have the comments of the authors on the actual tube life which has been obtained in the regulator circuit alone. It is understood that so far as tubes are concerned, the electronic regulators described have dual circuits and dual tubes so as to provide protection against tube failure. It would be of interest to have the authors comment on the effectiveness in their operating experience thus far of the dual tubes in preventing failure of the voltage regulating equipment. Also, have there been any failures which might be attributed to the use of dual tube circuits, as compared with the use of single tubes and associated circuits.

Summarizing the possibilities of various types of voltage regulators—there is still room for improvement in the simple electromechanical regulators. In the static network types some experience already has been obtained, but much remains to be secured. The electronic type regulator seems to afford many opportunities for development, not only in the regulating circuits, but also in associated controlling or limiting features. It is recognized however that a review of this situation several years hence may show many changes in the picture.

J. B. McClure (General Electric Company, Schenectady, N. Y.): The authors are to be commended for their excellent paper in which they review their operating experience with an unusual variety of excitation systems.

So far as exciters are concerned, probably the newest combination is the direct-driven overhung exciter operating in hydrogen. Direct-driven exciters long have been the preferred source of excitation for synchronous machines and first were abandoned for synchronous condensers with the advent of hydrogen cooling. Their return now marks a definite advance in design accomplishment and the maintenance and inspection record of these exciters bears out the fact that they are here to stay, at least for moderate speed machines.

Regarding the primary function of an excitation system which is to maintain output voltage, we have seen an excellent record in Figure 7 of the paper. It so happens that this record was made with an electronic exciter and regulator controlling. Experience indicates that comparable results will be obtained with modern excitation systems employing rotating exciters. It is now evident that an excitation system will be satisfactory for most applications if

1. The regulator is of the continuously acting type.
2. The regulator has a small dead band.
3. The exciter has at least moderate rate of response.

These requirements can be met with several combinations of equipment, and experience alone will indicate which combination is the most practical.

A. A. Johnson and **R. F. Lawrence** (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The authors of this paper and all participating operating companies are to be congratulated for making available to the industry the carefully obtained results of their experience with newer types of excitation systems. Indeed it is gratifying to know that all of the schemes which have been tried have given very satisfactory operation.

Of primary interest is the fact that electronic exciters have been proved by tests and actual operating experience to be a reliable excitation system for synchronous condensers and synchronous generators. These results have given a stimulus to the thinking of operating engineers and undoubtedly additional main electronic exciters will be applied. All excitation systems to date have certain features which are undesirable. However, any excitation system which can be serviced without having to shut down the main machine has a distinct advantage from an operating standpoint. An electronic exciter satisfies this requirement but has a higher initial cost and an apparently high maintenance cost because of tube replacements. The paper under discussion indicates that three power tubes and 4.4 firing tubes are required for each year of actual operation. It also states that further operation should produce a substantial improvement in the record for power tubes. We certainly agree that further improvement will be obtained because in one case the tubes already have operated an average of 36,389 hours, and also because ignitron tubes are guaranteed for three years. There is some question as to whether or not the cost of maintenance because of tubes (which should not exceed in the future approximately $600 per year) exceeds the actual system cost of outages required for maintenance of rotating types of excitation systems. This will be important on systems where there is a continuous need for the synchronous condenser. It is of more importance with regard to synchronous generators. Probably, as far as generators are concerned, any excitation system which reduces the outages of a base load generator by one per year can be justified economically. The authors indicate that no outages have been required for maintenance of the electronic exciters but that an average of three outages per year have been required for inspection and maintenance of the enclosed hydrogen-cooled d-c exciter.

It is believed that the authors of the paper have selected a fair comparison of power tube life by omitting the failures during the first year of service. However, we would like to explain the low average of 5,900 hours for three power tubes at New Carlisle. One tube was evidently defective and had to be replaced as a result of an excessive number of backfires. The life of the other two tubes was shortened because of improper cooling.

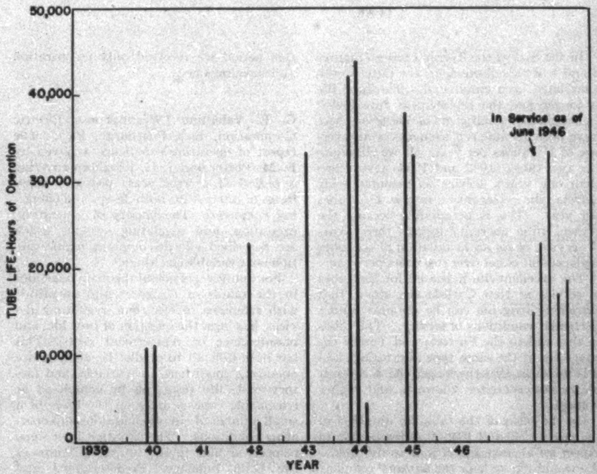


Figure 4. Ignitron power tube life at the Torrey exciter

The authors have given results of operating experience for electronic pilot exciters on condensers and turbogenerators and hydrogenerators. They also have given results of operating experience for electronic main exciters on four synchronous condensers. The discussers' company has built, tested, and placed in service a main electronic exciter for a 81,250 kva turbogenerator in the Springdale station of the West Penn Power Company. The results of these tests which were made in one of the a-c high power laboratories were presented in a recent paper.[1] This electronic excitation system is similar to that shown in Figure 4, of the paper, that is, the power supply to the rectifier is obtained from the generator terminals. Another main electronic exciter now is being built wherein the ignitron rectifier and associated control receives its power from an a-c generator directly connected to the shaft of the main generator.

It is not the intention of this discussion to justify any particular type of excitation system but to point out that the conclusions reached in the paper may be modified when the over-all system costs of the operation of each exciter scheme are considered. This is particularly important when considering excitation systems for generators.

REFERENCE

1. APPLICATION AND PERFORMANCE OF ELECTRONIC EXCITERS FOR LARGE A-C GENERATORS, H. A. P. Langstaff, H. R. Vaughan, R. F. Lawrence. AIEE TRANSACTIONS, volume 65, 1946, May section, pages 246–53.

E. L. Harder (Westinghouse Electric Corporation, East Pittsburgh, Pa.): At no time in the history of the industry has there been as great a variety of excitation and regulation systems from which to choose. This creates a problem for the prospective user in evaluating which system is best for his particular service and indicates a strong need for determination of the relative merits of the different schemes and stabilization upon a few of the best of them for general use. This is a period of transition from the older types of excitation and regulation systems to the newer electronic and rotating amplifier types with a great many conflicting requirements and tendencies. The final conclusions as to which systems are best can be reached only through appraisal of accurate information on the performance and characteristics of several available systems.

Thus, the paper of F. M. Porter and J. H. Kinghorn comes at a most propitious time. It is particularly significant in providing a mature evaluation of comparative merits and valuable field experience on a number of the excitation systems now undergoing trial.

The major stress has been laid on synchronous condensers, and several differences between the synchronous condenser and the turbogenerator or water wheel generator may be noted. In the case of the condenser, the continuity of the bus voltage is essential for the condenser to remain in synchronism regardless of its excitation system, and it therefore becomes more logical to take the excitation source from this bus provided the requirements of sufficient excitation during fault conditions can be met. This is generally possible as pointed out in the paper. On the other hand, in many turbogenerator installations, a 3-phase fault on the generator voltage system would lower the terminal voltage too low for it to be depended upon for excitation. In these cases, other sources of excitation power offer advantages. These are the main shaft, the station service supply, or a current compensator used in conjunction with the generator terminal voltage.

The static type of regulating network now called an impedance type regulator has received rather intensive development during the war period because of its advantages in naval service. It also offers certain advantages in central station service. It may be used to control either a Rototrol pilot exciter or a Rototrol main exciter.

The simplest of all rotating-amplifier-type regulating systems is the main exciter Rototrol, one of which is being built experimentally for a 60,000-kw turbogenerator. This offers the advantages of complete elimination of the pilot or control machine. The impedance type regulator or the electronic regulator directly controls the main exciter Rototrol. With the impedance type regulator there is, of course, no tube replacement at all. However the electronic regulator offers the advantages cited in the paper of adaptability to the addition of other control or limiting functions.

It is significant that the authors of this paper have concluded that the operation of the several devices, namely, electronic pilot exciter, electronic regulator, electronic main exciter, and rotating amplifier have been perfectly successful and that the decision as to what scheme should be used therefore hinges on initial and maintenance cost, and on the reliability of the power supply rather than on questions of reliability of the device itself. If this point of view is adopted generally, it greatly will facilitate reaching general industry conclusions as to the best ultimate form of excitation systems to be adopted for the different classes of service, namely, condensers, turbogenerators, and water wheel generators. It is to be hoped that within a few years, this situation will have stabilized so that in place of the large number of different types of excitation and regulation systems which must be considered now, a relatively few standard arrangements will be adopted by the industry. This paper should be of considerable help in clarifying this situation.

F. M. Porter and **J. H. Kinghorn**: H. C. Steiner's analysis of tube life of ignitrons is a welcome supplement to data in Table III of the paper, and gives a basis for optimism that further development work will improve the competitive position of electronic exciters in the field of synchronous machine excitation. In evaluating such a position, consideration must be given not only to maintenance costs of which tube life represents a major portion, but also to original cost. In addition, if savings result from reduced maintenance outages of the main generator or synchronous condenser, usually more important in the case of a generator, these must be included. While it is true that rectifier designs have been improved and standardized, their cost, including all necessary auxiliary equipment, is not yet as low as that of a direct-connected exciter. Manufacturer's warranty of tube life is a measure only of maximum replacement costs to the user and, as is suggested by Steiner's charts, it is hoped soon will represent a pessimistic evaluation of tube costs. In view of the variables involved, no general conclusion can be reached by the authors as to an average tube life which can be considered satisfactory.

If, as Steiner suggests, the excellent tube performance at New Carlisle can be attributed to lower exciter kilowatts and higher voltage, these factors should be considered in the design of electronically excited machines or in the selection of tubes for use with standard designs of machines. To state this in another way, the inference can be drawn that possibly tubes should be rated more conservatively when used with electronic exciters than when used for normal rectifier service.

It is interesting to note that C. E. Valentine concurs with Steiner in concluding that improved tube performance can be expected in the immediate future. Valentine's statement that a simple regulator is desirable because maintenance and trouble shooting then require less skill is one with which we believe most users will concur. This was a point which caused us some concern when we first began using electronic regulators. However an excellent and highly detailed set of trouble shooting instructions have been developed which have proved entirely satisfactory and have reduced maintenance to the routine level which exists on other better known equipment. So far as the use of dual tube circuits in the regulators is concerned, experience has not indicated any special advantage or disadvantage over single circuits. As tubes approach the end of life, their characteristics change gradually enough so that periodic testing and slow changes in regulator characteristics indicate this condition and new tubes are installed before actual failure occurs.

J. B. McClure's remarks about the requirements of an excitation system are not borne out entirely by any evidence presented in this paper. All of the regulators cited are of the continuous acting type and it is our experience that this is always preferable to a regulator of the dead-band type. A small dead band, if small enough, would approach a continuously acting type. Our experience with dead-band types indicates that the dead band cannot be reduced much below one per cent without specially designed equipment, and that a dead band of this magnitude is sometimes objectionable from the standpoint of voltage regulation and is usually objectionable from the standpoint of parallel operation with other regulators.

A. A. Johnson and R. F. Lawrence refer to statements in the paper indicating that no outages have been required for maintenance of electronic exciters, but that an average of three outages per year have been required for inspection and maintenance of the enclosed hydrogen-cooled rotating exciter. In the case of hydrogen-cooled condensers, it is necessary to open and clean the collector compartments three times a year, and therefore these outages occur on machines with electronic exciters as well as those with rotating exciters. In each case the outage can be charged more logically to the collector rings and hydrogen operation, because cleaning of an exciter compartment takes about the same time as cleaning of a collector compartment. Regarding operation of direct-connected exciters, we believe our experience indicates that maintenance for hydrogen-cooled exciters is going to be less than for air-cooled exciters. In order to keep brush wear and commutator wear to a minimum it is necessary to retain moisture in the gas, and therefore we have supervised closely the gas humidity. Although provisions have been made for adding moisture with the make-up gas, it has been determined that at least for five years the moisture originally in the machine is sufficient for satisfactory performance. This moisture apparently does not leak out of the shell appreciably with the leakage hydrogen.

Obviously, it cannot be expected that exciters can run indefinitely in hydrogen without inspection. It is our practice to make a general inspection of hydrogen-cooled condensers every five years, and the major exciter maintenance will be taken care of at these times. In the case of the Muncie condenser, although no trouble was experienced, the exciter was disassembled after three years operation for a check inspection. It was found that all parts were in excellent condition. Therefore we agree with McClure's remarks that hydrogen-cooled exciters for synchronous condensers are here to stay.

As to maintenance on direct-connected exciters of turbogenerators, we have found that in general necessary inspection and repairs of the exciter can be taken care of during the usual outages required for maintenance of the turbine and generator. Therefore, the total outage of the unit will be approximately the same regardless of the type of exciter.

E. L. Harder has mentioned the development of two very interesting additional methods of excitation, namely, a method which uses a current compensator in conjunction with generator terminal voltage as a power source and a method using a main exciter rotating amplifier. It will be very helpful if operating experience data can be published on these systems when available so that their performance and advantages can be judged in relation to the better known excitation systems.

As the discussion by Johnson and Lawrence indicates, all factors involved must be considered before a decision can be reached as to the proper excitation system for a specific application. In reaching such a decision it is very helpful to feel confident that all systems considered will operate reliably and accurately, and it is the authors' hope that the data presented will contribute to such confidence.

Influence of Magnetic Materials on the Welding Characteristics of Resistance Welding Machines

Discussion and author's closure of paper 46-137 by J. J. Riley and C. E. Smith, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, December section, pages 852–60.

A. C. Johnson (Progressive Welder Company, Detroit, Mich.): I would like to compliment the authors for having selected a subject which has presented for some time an annoying problem to those who are concerned with the design, application, and use of a-c resistance welding equipment. These authors have made a distinctly valuable contribution to the art.

I should like to add a word of caution concerning the accuracy of measurements to those who might continue the work that these authors have started. The paper states that the resistance between electrodes was measured oscillographically and "by taking the voltage drop across the two electrodes at two points as close to the weld as possible and dividing this voltage by the secondary current. . . . The inductive pickup is eliminated by carefully twisting the voltage leads together and holding them perpendicular to the electrodes." The inductive pickup by no means is eliminated by twisting the measuring circuit wires together. In fact, the induced voltage in the measuring circuit loop is quite appreciable. Naturally, the lower the resistance being measured, the more important is the effect of the induced voltage. The values of resistance between electrodes and resistance between sheets given in the paper vary from 10 to about 250 microhms, and, thus, the voltages measured range from approximately 0.1 to 2.5 volts. The voltage induced in the measuring circuit loop is of the order of 0.05 to 0.08 volt. From some of the measurements taken, the induced voltage is definitely negligible, especially as a result of the quadrature relationship between the two voltages under discussion. However, in other cases, especially those concerned with sheet to sheet resistances, the value of the induced voltage is considerable.

Assuming that the electrodes are part of a straight isolated conductor, the voltage induced in a rectangular loop is given by the equation[1]

$$c=\frac{3.19lw\,I_{\max}\cos wt}{2\pi 10^8}\times\left[\frac{1-3b^2}{4(1-b^2)}+\frac{b^4}{(1-b^2)^2}\log_e\frac{1}{b}+\log_e\frac{1}{c}\right]$$

$b={}^{r_0}/_{r_1}$, $c=s/r_1$

s = distance from the center of the conductor to the point at which the leads are twisted

r_1 = the radius of the conductor

r_0 = the inner radius of a hollow conductor (This is equal to zero in this case.)

1 = the points of attachment to the welder electrodes

It is also stated in the paper that the measurements of the resistive voltage component V_{rs} and the reactive voltage component V_{xs} were made using pointer stop meters. This method has an inherently low accuracy, and I believe that since these values were taken without any material actually between the dies, it would have been possible to use a longer duration of current flow and steady-state meter readings. Those values which are being determined, namely, the increase of resistance and reactance resulting from the addition of magnetic materials into the throat of the welder, are obtained of necessity by subtracting relatively large quantities. Consequently, even small errors in meter readings can produce considerable errors in the ultimate result. It is therefore advisable to use a technique of measurement in which every possible effort has been made to eliminate small errors.

I should like to point out also that the very difficulty with a-c welding equipment which makes this paper so valuable a contribution is non-existent in d-c welders. Therein lies one of the important advantages which welders having a d-c source of power supply have over the conventional type of a-c resistance welder.

REFERENCE

1. THE MEASUREMENT OF LARGE VARYING CURRENTS, A. C. Johnson. AIEE TRANSACTIONS, volume 65, 1946, January section, pages 8–11.

F. L. Brandt (nonmember; Thomson Electric Welder Company, Lynn, Mass.): The more I studied and the more I tried using the data contained in this paper, the more I realized how much work and thought have gone into its presentation. These data are much needed in the industry, and I wish this information to be common knowledge.

Formulas 1 and 2, Appendix I, use $Z_0 + Z_s$ arithmetically. I believe this should be a vector relation.

I find slight discrepancies between values obtained in Figure 12 in comparison with Figure 20. I am not able to find reference or derivation for the formulas for V_{xs} and V_{rs} given at the top of Figure 20.

C. M. Rhoades, Jr. (General Electric Company, Schenectady, N. Y.): This paper is very valuable in that it outlines for the first time a method by which the effect of magnetic material in the throat of resistance welding machines can be calculated with engineering accuracy by means of the empirical data presented in the paper. Heretofore, the characteristics of resistance welding machines could be determined only after a machine had been built and tested with the desired thickness and size of welding material in the throat. While it is true that most welder manufacturers have available information on the short-circuit kilovolt-amperes and power factor of welding machines, it has not been possible up to this time to determine accurately the welder demand and power factor under welding conditions.

As a result of the data given in this paper, it will be possible now for anyone who has short-circuit kilovolt-amperes and power factor data on a resistance welding machine to calculate accurately the conditions which will be obtained when material is in the throat of the machine. With this valuable data available, manufacturers of welders should be in a position to predict the actual conditions which will be obtained with production welding.

J. J. Riley: It has been pointed out by A. C. Johnson that a comparatively large error could exist in the sheet to sheet resistance measurements recorded in the paper. However, the actual welding resistance measurements shown in detail are of such magnitude that the error resulting from undesirable flux linkages is normally smaller than the error that could be expected from dirty, oily steel sheets or completely degreased steel sheets. Average production conditions formed the basis for the selection of laboratory conditions for measurement.

While not specifically mentioned in the paper, the impedance readings, in which meters were used, were taken with accurate meters in which temporary pointer stops were incorporated. However, the current flow time during any reading lasted between one and two seconds. The pointer stop was used to set the meter reading at approximately the correct scale position, but the meter actually reached steady state conditions before the current flow ended. It was desirable to limit the current flow time and consequently limit the amount of heat generated through eddy current losses in the steel sheet.

The paper does not explain the derivation of the formulas for V_{xs} and V_{rs}. The basis for such derivation, using the graphs contained in the paper, is explained in reference 10 of the paper.

Systems Development of Military Communications

Discussion and author's closure of paper 46-161 by Thomas R. Putnam, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 757–61.

Lee L. Glezen (Bell Telephone Laboratories Inc., New York, N. Y.): This paper is a real contribution to an understanding of one of the basic problems concerning communications for the Armed Forces. While dealing largely with the problems of organization in the North African Theater of Operations it confirms similar experiences occurring a year or more later in the European Theater of Operations. In the European Theater of Operations the widespread use of rehabilitated commercial cable, extensive Army-built open wire lines, spiral-four cables and radio formed a network of nearly 400,000 miles of main-line telephone circuits. This greatly exceeded any pre-World War II plans or expectations. These reached not only all Army and Air Force units on the continent but reached from Berlin, Munich, Bremen, and Frankfort to Paris, Marseilles, Cherbourg, LeHavre, Brussels, London, and even Washington, D. C. The number of long distance telephone calls handled for the Army and Air Forces at Paris alone reached 54,000 calls a day by June 1945.

The extensive technical problems of both wire and radio communications together with radar, television, and other special equipments following the technical principles of communication equipment are such that integration and team work are basic to assure satisfactory operation and performance. In World War II the cost of signal communications equipment for the ground and air forces reached the huge sum of about 5 billion dollars, exclusive of radar equipment which added over $1\frac{1}{2}$ billion more. These figures did not include the worth of rehabilitated communication plants liberated or captured and put to use to serve the Armed Forces in Africa, France, Belgium, Luxemburg, Germany, and many other places. The maximum strength of the Signal Corps maintained throughout most of the war period was about 370,000. This was exclusive of the Navy's requirements. Certainly, for such a huge task a well-established and properly functioning organization is required to provide economically and efficiently the team work required for full integration of communications.

Peace time consideration of communications and organization for any possible future conflict should avoid basing plans too much upon what was found to be the need for World War II, just as it was found that World War I needs were outmoded for World War II. In other words, a carefully integrated communications organization is required, but it should be left sufficiently flexible to encompass any developments yet to be conceived and complete enough to assure team work from top to bottom of the communications field as new ideas are injected in the future.

The author is to be congratulated for bringing this matter before the Institute at this time.

C. O. Bickelhaupt (American Telephone and Telegraph Company, New York, N. Y.): The author refers to the fact that "in those overseas theaters in which organizational difficulties were encountered in systems development of communications during World War II, these difficulties stemmed from a basic deficiency in signal communications doctrine as promulgated by the War Department." He points out further that "this doctrine sets forth the principle that every commander controls the system of communications required for his command."

The doctrine referred to perhaps is stated better by saying that "communications is a function of command." The difficulty to which Colonel Putnam refers arises not from the fact that this doctrine is wrong but rather comes through misunderstanding of the doctrine and from its improper application.

On the one hand, the chief signal officer of a theater or subordinate command is primarily a staff officer and as such ordinarily does not exercise command. On the other hand, the chief signal officer does have staff responsibility for the adequacy, co-ordination, and satisfactory operation of signal communications within a given command. In fulfilling this obligation he issues technical directives through technical or staff channels and initiates command directives for issuance by his commander. The command channel, however, flows past him from the commander to the next subordinate commanders.

Apparently in the North African Theater of Operations, the American Army ran into this difficulty, because Colonel Putnam states that "it was soon found that staff responsibility without technical administrative authority over the operating personnel of the various subordinate commands was entirely unsatisfactory." This difficulty was solved apparently through staff channels rather than command channels by giving "technical administrative authority" to the chief signal officer of the theater and echelons of his organization. Apparently this worked out satisfactorily.

In the European Theater of Operations, with which I am familiar, this problem, however, was approached from what I consider to be a more fundamental basis, and its solution was reached in the command channel rather than in the staff channel. This was done

1. By establishing an independent command known as the "military communications service" and commanded by a director general, military communications service.

2. The director general, military communications service, was set up to report directly to the superior commanding general, and the military communications service, therefore, received its command authority directly from the superior commander. The director general, military communications service, was made co-ordinate in command channels with area commanders throughout the theater.

3. In order to ensure proper staff and technical direction and co-ordination, the superior commander gave technical operating control of the military communications service to his chief signal officer.

Thus, while the military communications service received directives as to its mission and as to technical requirements directly from the chief signal officer who, in issuing such directives, was speaking with the voice of the superior commander, the director general, military communications service, was in his own right an independent commander and could command his troops, assigning their specific missions within his general directives, moving them from one geographical command to another, and having complete responsibility for results.

This solution is not new and was, I believe, first used in the American Army in the construction and operation of military railroads where the military railway service was established as an independent command receiving its command authority from the theater commander but operating under directives of the chief of transportation, a staff officer on the staff of the theater commander.

Colonel Putnam is absolutely correct in his conclusion that in order to secure systems operation of a military communications service command authority is essential.

Thomas R. Putnam: General Bickelhaupt has brought out two important points very ably in his discussion. First, he takes the view that the War Department doctrine under which signal communications operations were carried out in World War II was not deficient but was misunderstood and applied improperly. I am not inclined to quarrel with this view, for, in either case, attempts to apply organization principles looking toward system operation of communications in the field encountered difficulties, as General Bickelhaupt agrees.

The second point which General Bickelhaupt discusses is the question of command versus staff channels for the exercise of technical administrative authority in matters relating to the operation of military communication systems. In my opinion there is no question but that the solution arrived at in the European Theater, establishing the military communications service as a "line" organization for the operation of the communication system, was far superior to the arrangement in French North Africa described in the paper under discussion. In the North African Theater of Operations there were numerous factors considered by those responsible for the final decision in this matter, one of which, for example, involved the geographical distribution of Allied communications troops in the area served by the land line system. For this reason, the North African arrangement, which established a certain amount of staff authority for technical administration purposes, was a compromise solution.

The military communications service in the European Theater, referred to by General Bickelhaupt, was not established until July 1945, thirteen months after the Normandy invasion and two months after V-E Day, despite earlier efforts of those charged with the responsibility of operating the communication system. Likewise, in French North Africa suitable centralized control of the land line system did not take place until September 1943, which was ten months after the Allied landings and four months after the close of the campaign in Tunisia.

Glezen in his discussion has given us some interesting data concerning the magnitude of the military signal communications operations in Europe, as well as an indication of the total communications resources required by the ground and air forces in the prosecution of World War II.

Concerning Glezen's ideas as to the flexibility of basic principles of military communications organization, I am in complete agreement.

A 400-Ampere Sealed Ignitron

Discussion and author's closure of paper 46-140 by H. C. Steiner and H. N. Price, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 680–5.

W. E. Pakala (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The authors have given a good explanation of the several factors influencing arc-back and how they may affect the design of mercury-arc power tubes. The effect on arc-back rate and the relative importance of all these factors is not known at the present time, and the design of ignitrons as well as other types of mercury pool rectifiers is based, therefore, largely on our knowledge of design factors of tubes which have proved satisfactory in service.

About four years ago, when we began the development of a 400-ampere sealed ignitron, it was decided to base the sealed design on pumped ignitrons which performed so well in the electrochemical and other fields. The new 8-inch diameter 400-ampere sealed ignitron design proved entirely satisfactory on test. Several sealed tubes were placed on life test for 14 months to determine vacuum tightness of the arc-weld construction. Only two tubes out of a total of ten were found to be gassy. (With the present system of mass spectrometer leak detection used in the shop, all tubes will be vacuum tight.) The others performed perfectly in a 6-phase rectifier circuit under normal load, overload, short circuit, and under other test conditions necessary for the development of a new tube design.

In Figure 3 of the paper by Steiner and Price, are shown oscillograms indicating the deionization rate. These oscillograms were obtained with a 100-ampere load in each case. From the data, two interesting conditions are indicated. These are

1. The deionization time constant increases approximately as the second power of the tube diameter. (The data apply only to tubes without grids.)

2. The cooling water temperature has the same effect on deionization as changing the tube diameter. Figure 3 shows that for constant current the deionization at 20 degrees centigrade for the 8-inch tube is about the same as deionization at 60 degrees centigrade for the 5-inch tube.

The data discussed above indicate that this arc-back influence factor increases with the tube diameter. Therefore, the volume density of ionization and the current density per unit area of tube cross section is reduced considerably in the design of larger tubes. The net result is that the full load current rating is increased for tubes of similar design only as the first power of the tube diameter. For short-time rectifier ratings, such as the 200- and 300-per cent overload ratings, the performance improves with increased tube size because the heat capacity increases more rapidly with increase in the tube diameter than the current rating.

As to the maintenance of vacuum, the paper points out that the arc and the walls of the tube have a gas cleanup action. This cleanup or pumping action is the reason for reduction in life of gas discharge tubes of various kinds. In the case of most mercury-vapor tubes, however, this action does not reduce the life but actually increases it. In fact high fault current tubes, such as high-power ignitrons, could not be operated at all without pumps unless the arc-back rate were practically zero or very fast circuit breakers were used. This is so because of the gases given off by various tube parts during high current arc-backs. These gases are removed rapidly by the pumping action of well degassed walls, and no steady-state pressure will result to affect tube performance.

The experience in this country with large sealed ignitrons and in Europe with sealed metal multianode type rectifiers indicates that vacuum-tight tubes of large size can be built by using arc or gas welding or seam welding for closing headers to the main cylinder of the tube.

H. C. Steiner: W. E. Pakala's comments and particularily his discussion of the effects of tube size and cooling water temperature on deionization rates are much appreciated.

Our experience indicates also that overload capacity is improved with the increased heat capacity of larger tubes.

The gas cleanup action of the walls and arc is important in maintaining a high vacuum in sealed tubes. High speed anode circuit breakers usually are employed to protect both the equipment and rectifying elements, whether sealed or pumped, when fault currents become high. In the case of smaller rectifiers, such as those using 100- and 200-ampere size sealed tubes, medium or slow speed circuit breakers (three to eight cycles) are used satisfactorily. One factor, as Pakala points out, permitting the use of these relatively slow circuit breakers is the rapidity with which any gas is removed; another is the extremely low rate of arc-back in the sealed tubes.

A New Design for the A-C Network Analyzer

Discussion and authors' closure of paper 46-135 by J. D. Ryder and W. B. Boast, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, October section, pages 674–80.

Eric T. B. Gross (Illinois Institute of Technology, Chicago, Ill.): Amongst other features, it is a timesaving advantage of this analyzer that the voltage of each generator is controlled separately by a voltage regulator. The practical problems to be solved often require that the internal voltage, the terminal voltage, or perhaps the bus-bar voltage be kept constant. In load studies bus-bar voltages have to be set and kept at a fixed value; in short-circuit and stability studies the internal voltages are unchanged.

In this analyzer, provision is made for the internal voltages to be controlled; it would be helpful to arrange a switch on the panel of the generator unit so that the voltage control could be taken from points not only before but also behind the generator impedance or, in case of unit-connection, from the high-voltage bus-bar.

Abe Tilles (Pacific Gas and Electric Company, San Francisco, Calif.): The authors obviously have carried their original development work a long way and have produced a very neat and interesting instrument.

Errors caused by stray capacitance from leads and relays are one of the limiting factors of the lower-frequency conventional network analyzers. Because of the higher frequency used here, together with fairly low impedance and power levels, it seems certain that capacitance errors will be definitely a limiting factor in the use of this analyzer. The use of concentric cables further increases the lead capacitance.

It is therefore of interest to inquire for additional detail on what steps were taken to reduce stray capacitance; what amounts were tolerated; and how they affected operation of the equipment.

The matter of stating a fair test, usable in a commercial specification, to measure simply and definitely the degree of troublesomeness of stray impedance and admittance and of induction between circuits is one which is quite unsettled and difficult even for the 400-to-500-cycle range of present commercial equipment. Therefore, it is of interest to know with what tests or requirements the authors approached this question in regard to their analyzer. What do they consider a fair measure of these items?

C. Concordia (General Electric Company, Schenectady, N. Y.): In regard to the generator units, I should like to ask the authors whether they have data similar to that of Figure 6 of the paper for the phase angle regulation.

Lester LeVesconte (Illinois Institute of Technology, Chicago, Ill.): The authors should be congratulated on the courage to pioneer any development such as they have undertaken and on the success they have achieved. They should be commended on their general design, such as the choice of the frequency, base currents and voltages, circuit selector system, and metering equipment. While suggestions could be made or words of caution could be offered on many details, these remarks will be limited to a few of the most important items, based on operating experience at the Armour Research Foundation of the Illinois Institute of Technology in Chicago.

The paper mentions that the power dissipated in the network is so low that small resistors may be used safely without fear of burn-out on accidental overloads. The actual limit or current rating is not given but would be of interest. Of the two designs in most general use, one has a limit of about three times the base current, and the other has a limit of about ten times the base current. The lower limit has been found to be a handicap in some cases, and, therefore, a higher current rating is desirable if it can be obtained at a reasonable cost.

It seems that the length of the connecting leads could be a limiting factor, especially since the longest lead on the present pilot model seems to approach the permissible limits of capacitance. It must be recognized that these leads will become necessarily much longer as the equipment is expanded to the size required for most present-day power system problems. However, this difficulty would disappear in an assembly made to represent only a specific power system, using fixed values instead of adjustable units and eliminating the plug and jack panel. It would seem that the design developed by the authors, using standard parts, would be adapted particularly well to this field, and this might make individual analyzers practical for many more operating companies.

On the meter desk, all of the controls are located on the left to be operated by the left hand, for the purpose of leaving the right hand free to record data. Actually, the location of the selector push buttons on the right does not interfere with the recording, but aside from this there are good economic reasons for designing the equipment for at least two-man operation, having one person to read the instruments and another record the data. The operation of the modern analyzer represents a cost of from $5.00 to

Figure 1

$20.00 per hour, depending on the number of people involved, thus making any time-saving equipment or methods very valuable. Using the same general type of circuit selector described by the authors, a new recording table designed and built at the Armour Research Foundation has been found to be particularly useful. (See Figures 1 and 2 of this discussion.) It is located adjacent to the meter desk and connected to it by a multiple cable. The sloping top has a glass panel on which the system diagram is placed and beneath which are small lamps which indicate the circuit being metered and thus show the location on the diagram where the readings should be recorded. The lamps are spaced one inch apart in each direction, so that a light position can be associated with each circuit element on the diagram. Each lamp is connected to a small jack on a panel on the back of the table. Below the jacks is a set of plugs on retractable cords, each numbered to correspond with a circuit on the analyzer. Since any plug, representing a circuit element on the analyzer, can be inserted into any jack, it is possible to assign whatever circuit number is desired to each lamp position beneath the diagram.

The plugs are energized by metering relays which are operated by a circuit selector controlled by a set of push buttons on the meter desk. These relays are located back of each circuit element on the analyzer, and only one such relay can be energized at a time. Besides connecting that circuit to the meter bus, one contact of the relay energizes a light on the circuit drawer, indicating which circuit is being metered. The plugs are connected so that the lights on the table operate in parallel with the lights on the circuit drawers. Thus, as each circuit in turn is set up on the push buttons at the meter desk, the circuit selector energizes the proper relay to connect that circuit to the meter bus, and the same relay energizes the light on the circuit drawer and through the plug and jack assembly energizes the light beneath the proper circuit on the diagram.

On the meter desk is a switch which is marked + and − and which reverses the direction of the watt-varmeter. The position of this switch when the meter reads upscale is an indication of the direction of power in the circuit being metered. An extra contact on this switch controls the direction-indicating lights on the recording table. The + light is located in the upper, left-hand corner of the table, and the − light in the upper, right-hand corner, so that the person recording data can see easily which is lighted. The recorder has time to observe and mark the direction while the reading is being made.

Experience has indicated that a few people who have not seen or used the table have questioned its value, but everyone who has seen it has been convinced that it is a great timesaver and definite contribution to the speed and accuracy of operation. It results in all the data being recorded directly and quickly on the system diagram. It eliminates the possibility of recording readings on the wrong circuit. It provides a safeguard against reading the wrong scale on the meter or using the wrong multiplier, because the values around each junction point can be checked easily. It makes practical the recording of only those values that are significant in cases where a minor change affects only a limited area of the system, thus eliminating a volume of unnecessary readings.

From Figure 7 of the authors' paper it appears that the instruments on the meter panel for each generator do not have normal indexes or fixed pointers. These pointers should be added. They have been found to be very helpful in adjusting the generators in the original setup and also on any subsequent system changes. This same idea has been extended to the adjustment of the load circuits at Armour Research Foundation in the manner shown in Figures 3 and 4 of this discussion.

It will be noticed that the connections shown are fundamentally the same as the load circuits described by the authors in their Figure 10, with the addition of a small voltmeter that indicates the voltage across the load impedances beyond the load adjuster. Each voltmeter is mounted directly beneath its associated load adjuster and has a scale reading from 80 to 120 volts. It is provided with a normal index that is set to indicate the voltage after the load has been adjusted by test to give the correct value in watts and vars. For any subsequent system changes that might affect the voltage at that point on the system, the load can be corrected by means of the load adjuster simply by bringing the voltage back to the original value, as indicated by this small voltmeter. For most system changes dur-

Figure 3 (right)

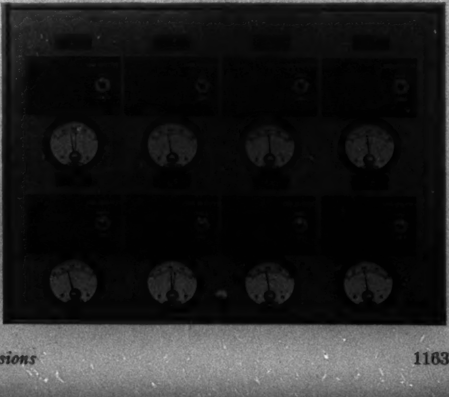
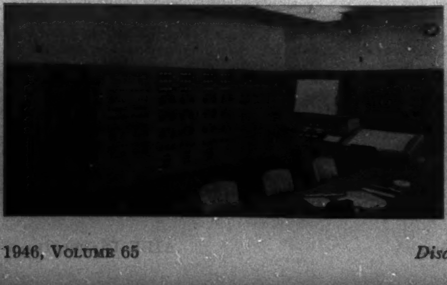


Figure 2 (below)



ing the study, most of the readjustments can be made by means of the generator and load circuit instruments, thus saving considerable time.

H. L. Clark (nonmember; General Electric Company, Schenectady, N. Y.): This paper illustrates the present trend toward use of electronic circuits to extend the range of usefulness of a-c network analyzers.

It would seem that emphasis in this paper should be placed upon the convenience and speed with which the correct answers to problems can be obtained with this network analyzer as well as reduction of first cost, because if the analyzer has a reasonable degree of utilization, it is the operating cost per problem which is of importance rather than the initial cost of the analyzer.

Additional information on the master instrument system would have been of interest since many of the problems are concerned with the instrumentation, especially when operating at a frequency of ten kilocycles. For example, the burden of the instrumentation and the accuracy achieved in phase angle measurements should have been given.

A considerable point was made concerning the accuracy and stability of the frequency source. It is implied that the temperature-controlled quartz-crystal oscillator held to 0.001 per cent by comparison with the WWV transmissions is an essential part of the equipment. Since the reactances vary only as the first power of the frequency, and since these units are adjusted to 0.2 per cent accuracy, it is hardly necessary that the accuracy of the frequency be 200 times better.

The paper states that the capacitance introduced to the metering bus by the contacts of the switching relays has been compensated effectively and the resultant error made negligible. Since compensation for capacitance is accomplished by the use of shunt reactors, it would be necessary to have as many reactors as there are contacts in the switching relays, which is, of course, out of the question for a network analyzer having a large number of units. Perhaps the author intended to state that this effect is reduced by sectionalizing the metering bus so that only a small number of switching relays are connected to the master instruments at any one time.

The presence of harmonics in the generator output can cause considerable error, especially in cases where due to a special circuit arrangement the fundamental frequency current is cancelled out leaving only the harmonic currents. The master instruments should, in this case, indicate essentially zero current but will not do so unless the harmonic currents are absent or the instruments do not respond to these harmonic currents.

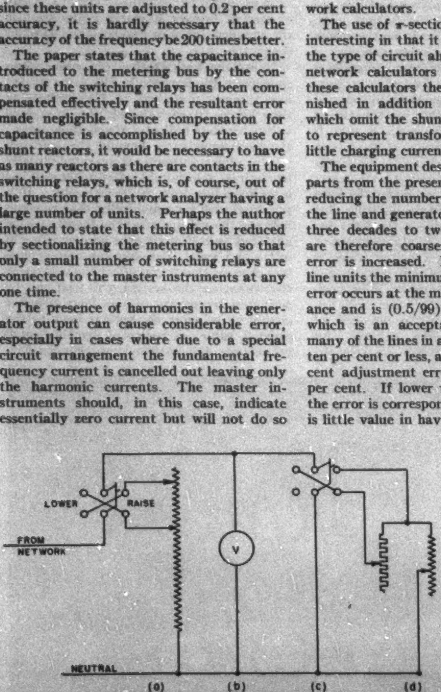


Figure 4

Because the type of generators used develop even harmonics as well as odd, it becomes especially important that attention be paid this detail. It would be expected that the presence of harmonics would also affect the phase angle measurements, since the type of circuit used measures with respect to the point the wave goes through zero. Harmonics of certain phase relationships will shift the point at which the wave goes through zero and hence cause errors in the phase angle measurements.

R. B. Squires (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The authors are to be congratulated for a good paper on an interesting development. They have solved most of the many problems inherent in designing a network analyzer or calculator to operate at relatively high frequencies.

The equipment contains several features which will increase the speed and ease of operation. The use of a few push buttons to control the metering relays of all the circuits permits a reduction in the time to meter the network. The use of load-adjusting autotransformers, three meters for each generator, and a separate instrument desk increases the flexibility of the equipment. The advantages of all these features have been proved time and again during the last ten years in the operation of many network calculators.

The use of π-sections for the line units is interesting in that it shows a preference for the type of circuit already furnished in four network calculators now being built. In these calculators the pi-line units are furnished in addition to standard line-units which omit the shunt branch and are used to represent transformers or lines having little charging current.

The equipment described in this paper departs from the present accepted practice by reducing the number of adjustment taps on the line and generator reactance units from three decades to two decades. The steps are therefore coarser and the adjustment error is increased. For instance, with the line units the minimum per cent adjustment error occurs at the maximum value of reactance and is $(0.5/99)\times 100 = 0.505$ per cent, which is an acceptable value. However, many of the lines in a problem may be set at ten per cent or less, and at this point the per cent adjustment error is $(0.5/10)\times 100 = 5$ per cent. If lower values must be set up, the error is correspondingly greater. There is little value in having circuits values held to an accuracy of 0.5 per cent or better if the actual error is five per cent in representing an impedance of 10.5 per cent. This limitation has been overcome in network calculators with which I am familiar by using three decades of adjustment or by using a calibrated, continuously variable reactor for the small values.

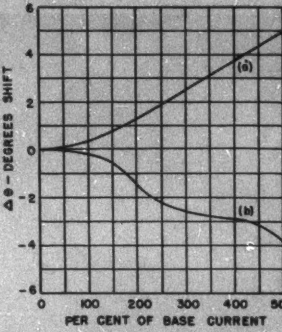


Figure 5. Phase angle regulation of the generator voltage for varying output

(*a*)—Current leading voltage by 90 degrees
(*b*)—Current lagging voltage by 90 degrees

Another way of stating the problem is that using only two decades of adjustment limits the range of settings to the higher values if good accuracy is to be maintained. Because there is wide variation in a given system in the values of impedances that must be represented, it is important to give each circuit a range as wide as possible over which accurate settings can be made. The use of three decades or a continuously variable unit increases the range in which accurate settings can be made.

It should be realized that the curve of Figure 1 of the paper applies only to capacitive reactance, which accounts for only a small part of the total cost. The cost of cabinets, instruments, plugging panels and the labor of assembly and wiring make up a large part of the total cost. The cost of these items stays constant or even increases with frequency because of the necessity for shielding and for more complicated wiring at the higher frequencies.

In connection with the instruments in the control desk, it would be interesting to know the current and voltage ranges used and the phase angle accuracy of the wattmeter and phase meters, since phase angle accuracy is as important to the over-all accuracy as the accuracy of calibration of the instruments.

Is it possible to measure directly the voltage drop across a circuit and the power loss in it? Such information is very desirable in transient stability studies in obtaining equivalents of networks and in other problems where the voltage between two points in the network is needed, rather than the voltage to neutral.

One of the problems in the design of a mutual transformer is to keep the capacitance between windings low, because in circuits where these transformers are used the voltage between windings is several times

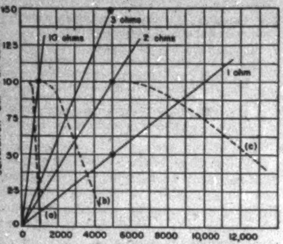


Figure 6. Design co-ordination of resistor ratings with generator volt-ampere characteristics under extreme overloads

the voltage across each winding. In most studies involving mutual transformers several transformers must be used, and the effect of the capacitance leakage currents is additive for each transformer. This problem must be particularly difficult at 10,000 cycles, and, therefore, it would be interesting to hear to what value the capacitance current was limited.

The discussion of these problems is not intended to minimize the good work the authors have done in producing a workable design at frequencies far above those normally used for measurement work. No doubt operating experience of the next year or so will show whether circuits of wider range are needed and whether any difficulties develop from the use of a large number of electron tubes.

J. D. Ryder and W. B. Boast: We wish to thank those contributing to the discussion. The answers to their questions are given in an itemized manner.

Phase Angle Regulation of Generators

Figure 5 of this discussion gives the phase-angle regulation curve of a generator for the extreme conditions of capacitive (curve *a*) and inductive (curve *b*) loading. Other loading conditions lie between these curves. The shift in phase of the voltage is caused principally by the plate resistance of the output stage of the amplifier. This resistance is reduced effectively by the use of triode output tubes and by the high degree of degenerative feedback employed. At 200 per cent current (the limit of the metering system) the shift in generator voltage phase is less than 1.5 degrees.

Voltage Regulator Control

The addition of a switch on the panel of the generator unit for providing voltage control on either the internal or external voltage of the generator is being made as a desirable improvement.

Capacitance Errors

A. *Relay Contacts Between Units and Meter Bus.* The metering bus was sectionalized into five regions (one for each of the major racks of the analyzer), and the capacitance to ground for each of these sections was measured. Inductive compensation then was established at the capacitive load center of each section. The metering bus then was connected electrically together again as a unit and operates this way in the analyzer. The compensation at relatively few points for the highly distributed nature of the relay positions has reduced effectively the relay capacitance current errors to approximately one-half per cent of base current when all elements on the network operate at the general level of base voltage. In all arrangements of relay contacts, those permanently connected to points in the high side of the network units proper have been placed with as much clearance as possible to ground and to other relay contacts.

B. *Connecting Leads.* The maximum length of leads between units and terminal board produces capacitive current errors within the allowable tolerance. These errors are stated in the original paper. A network analyzer with four to six times as many units as this pilot model could be built in sections, each essentially as the present analyzer. In setting up problems on this larger analyzer, relatively few long trunk leads between these sections would be sufficient. These trunk leads can be designed within the maximum tolerance on series resistance and reactance. If the shunt susceptance is too great, the same plan of compensation as is discussed previously in *A* could be applied to each lead. Therefore a larger network analyzer is not impossible at this frequency.

Burden of Instrumentation

The burden of instrumentation is stated in the original paper in the section entitled "Control Desk."

Frequency Accuracy

An accuracy of 0.001 per cent in frequency was not intended to be understood as an essential condition of accuracy. This frequency source was used because it was commercially available at a low cost.

Power Limit of Resistors

Figure 6 of this discussion shows graphically the conditions which can be encountered on accidental overloads through line units. The dashed curve (*a*) shows the extension of Figure 6 of the original paper to a much different scale. All of the information of the original Figure 6 is contained in the first one-half division on the "per cent of base current" scale. Beyond this point the voltage regulator loses control and the voltage drops to essentially zero at about 1,200 per cent of base current. Curve *b* shows the volt–ampere characteristic of four generators (as in the present pilot model) all operating in parallel on the same bus with their voltages all in phase. Curve *c* shows the same for sixteen generators under the identical condition of in-phase operation, a highly improbable condition but, nevertheless, the worst one in event that an accidental overload occurs on a line unit just past that bus. The line unit resistors in the lowest decade (1-ohm steps) are 25-watt units. The circle on the 1-ohm resistance line represents this power rating. If more than six or seven generators should be paralleled and act on this single 1-ohm resistor as a load, it would be in danger of excessive heating. If, however, two of these resistors were in the circuit (the 2-per cent step) the power rating of the two becomes 50 watts and the circle on the 2-ohm line shows that ten volts could be maintained by 16 generators all acting in parallel and in phase, giving exactly 50 watts dissipated with no danger of overheating. The 3-ohm point lies above the generator characteristics, and, hence, neither it nor any of the other settings to nine ohms could cause overheating. The next decade begins at ten ohms, and each of its resistors is rated at ten watts. The circle shows that no overheating would occur on it as base voltage.

The instrument amplifiers likewise possess a self-protective feature in their design, and, even though many times their normal full scale value is applied to the amplifier inputs, the actual instrument currents do not exceed 150 to 200 per cent of full scale current. Also, no damage occurs to the generators under these accidental overloads because the average plate power dissipated in the output stage is essentially independent of 10,000-cycle loading. The generators are fused however at 1,000 per cent of base current.

Phase Meter Accuracy

The accuracy of the phase meter is dependent upon the range of amplitudes for which the phase is to be measured. For currents from 200 per cent down to 15 per cent or for voltages from 150 per cent down to 10 per cent the accuracy is plus or minus two degrees. For smaller currents or voltages the accuracy becomes progressively poorer. However, high accuracy of angle determination of a small vector among large vectors is usually not possible by analytic methods.

Wave-Form Distortion

Measurable wave-form distortion occurs only as the voltage of the generators is driven above 125 per cent base value or as the current reaches 300 to 500 per cent of base value. The first of these conditions usually can be avoided and the second is impossible to meter.

Design of Induction Heating Coils for Cylindrical Magnetic Loads

Discussion and authors' closure of paper 46-124 by J. T. Vaughan and J. W. Williamson, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 887–92.

N. R. Stansel (General Electric Company, Schenectady, N. Y.): This paper including the companion paper[1] mentioned in the synopsis has two values. The first is its significance to those concerned with the art of induction heating. The second, of more general interest, is its worth as a study of the principle of electromagnetic induction.

It is well worth while to re-examine periodically the principle of the eddy current phenomenon. New conditions arise from time to time. The example here is the use of eddy currents for heating service. This viewpoint emphasizes the importance of considering the relations of the thermal properties of a material to its electrical and magnetic properties.

The initial procedure in such an examination is not concerned with whether the object is the minimizing of the eddy-current

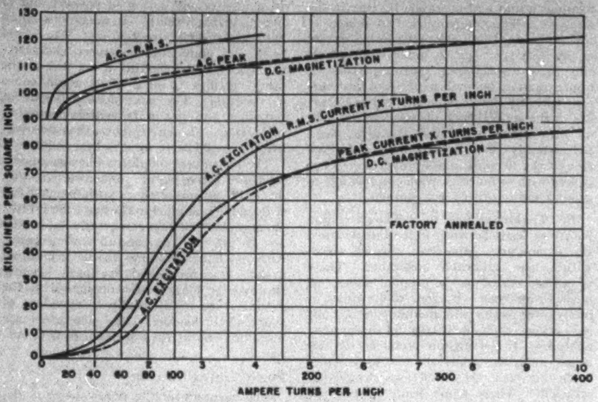


Figure 1. A-c excitation curve (60 cycles) and d-c magnetization curve of SAE1015 steel

Lines per square inch ÷ 6.5416 = lines per square centimeter

Ampere turns per inch × 0.4947 = oersteds

watts or the use of eddy currents for heating service. The basic relation is

$$P=(H^2\sqrt{\mu f}\sqrt{\rho 10^9})lS\times 10^{-7} \text{ watts} \quad (1)$$

This equation expresses the rate of heat development by eddy currents in a conducting mass which is exposed to a uniform and sinusoidal alternating magnetic flux, the directions of which are parallel to the lateral surface of the mass.

The symbols are

H = the magnetizing force (crest value), oersteds

μ = the a-c permeability of the material*

f = the frequency, cycles per second

ρ = the resistivity of the material, ohm-centimeter cube*

l = the length of the mass, centimeters

S = the shape factor of the mass—a matter of geometry only

For the solid cylindrical mass, or load, of the paper

$$S=\frac{aF_r}{4\sqrt{2}} \quad (2)$$

where

a = radius, centimeters

F_r = the function

$$\left[\frac{ber\ \Delta\ ber'\ \Delta+bei\ \Delta\ bei'\ \Delta}{(ber\ \Delta)^2+(bei\ \Delta)^2}\right] \quad (3)$$

Δ = the parameter (index ratio)

$$=\left(\frac{2\pi\sqrt{2}\sqrt{\mu f}a}{\sqrt{\rho 10^9}}\right) \quad (4)$$

Permeability. The a-c excitation curve (crest values) of a magnetic material more or less coincides with the d-c magnetization curve as the saturation point is approached. An example is given in Figure 1 of this discussion. Hence, in induction heating practice, the d-c magnetization curve—usually more readily available—is generally of sufficient accuracy. The a-c excitation curve should be used for lower flux densities.

REFERENCE

1. DESIGN OF INDUCTION-HEATING COILS FOR CYLINDRICAL NONMAGNETIC LOADS, J. T. Vaughan, J. W. Williamson. AIEE TRANSACTIONS, volume 64, 1945, August section, pages 587–92.

J. T. Vaughan and **J. W. Williamson:** Textbooks frequently develop the theory of the polyphase induction motor in terms of an equivalent circuit derived from the transformer. The analogies existing between these two pieces of equipment enable the student to gain a better understanding of both. If these analogies were pushed too far however, for example, if an attempt was made to adapt the equivalent circuit theory to the effect of skewing of the rotor conductors on starting characteristics, the resulting comparisons might become more of a hindrance than a help.

In a similar way, the concept of a constant permeability may be useful educationally in bridging the gap between the induction heating of nonmagnetic materials and that of magnetic materials. To avoid reaching incorrect conclusions, the following consequences of the fact that the B-H curve is not a straight line should be noted:

1. Theoretical formulas involving permeability require laboratory experimental evidence to establish the value of permeability which should be used, even if the B-H curve is known.

2. As noted in our paper and pointed out by N. R. Stansel, the conclusion reached by assuming any constant value of permeability that for moderately large diameter loads the load power factor is 0.707 is false. Actually the ratio of watts to volt-amperes of a carbon steel load is about 0.84, and the ratio of reactive volt-amperes to watts is not 1.000 but about 0.65.

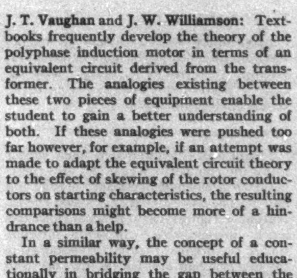


Figure 2. Transverse and axial flux

The discussion of transverse flux will perhaps be clarified by Figure 2 of this discussion. The number of lines of transverse flux at A or A' may be defined as the number of lines of flux which flow radially outward through the infinitely long cylindrical surface $AA'B'B$. Because lines of flux are continuous, the same number cut axially through the circular surface AA'. Therefore, it is immaterial whether one considers transverse flux or axial flux. The same current density i would be determined from equation 11 of our paper in either case.

Protection of Industrial Plants Against Insulation Breakdown and Consequential Damages

Discussion of paper 46-122 by H. R. Vaughan, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, August–September section, pages 592–6.

A. A. Johnson (Westinghouse Electric Corporation, East Pittsburgh, Pa.): Particular attention is called to Figure 8C of the paper. This scheme of high resistance grounding can be used readily on industrial distribution systems of 2,400 volts and above where no neutral is available for a grounding device. This relatively high resistance grounding device allows sufficient in-phase current to flow during the time of a line-to-ground fault to dampen possible high frequency voltage oscillations and thus to prevent excessive transient overvoltages. The reason for this may be explained briefly. For an ungrounded system if a line-to-ground fault appears the current which flows is caused by the inherent distributed capacitance of the system. This current leads the voltage by 90 degrees. Because of the fact that on most industrial systems this capacitive current is only a few amperes (three to ten amperes are common values), it probably will cease to flow at current zero. The current zero, however, occurs at a time corresponding to the peak of the voltage wave and when this current ceases to flow there is a tendency for the instantaneous voltage to be double normal peak line-to-ground voltage. Restriking of the fault arc at the optimum time relationship may produce higher peak voltages by a buildup process which may be several times the normal peak voltage to ground. This high peak transient voltage may cause failure of connected rotating machines, cables, and lightning arresters. By using the high resistance grounding device two main effects reduce transient overvoltages.

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* The values μ and ρ correspond to the mean temperature of the path of the current in the mass.

1. The fault current zero does not occur at 90 degrees with respect to the voltage but occurs at a

point probably less than 45 degrees with respect to the voltage where the instantaneous voltage is considerably less than the peak value. Since the high frequency oscillation takes place about the point on the voltage wave where the current goes through zero, voltages which do occur because of restriking therefore are reduced very greatly. When the loss in kilowatts in the resistor is made equal to the distributed capacitive kilovolt-amperes to ground of the system, the resistor will limit the maximum peak line-to-ground voltage to about 2.5 times the normal peak line-to-ground voltage.

2. Reducing the time constant of the fault circuit damps the high frequency oscillation.

There are a number of systems, both industrial distribution and utility distribution, where failures of electrical equipment caused by transient overvoltages can be reduced materially, if not eliminated, by the proper application of high resistance grounding. Simultaneous ground faults on an ungrounded system are strong evidence that transient overvoltages are causing failures. By the use of a resistance grounding device the line-to-ground fault current is increased approximately 50 per cent above what it would be for ungrounded operation. Thus the system still can be operated as an ungrounded system but with the one serious disadvantage of ungrounded operation minimized to a point where little or no trouble should be expected from excessive transient overvoltages. This scheme of high resistance grounding is now in service on several distribution systems.

Electromechanical Transient Performance of Induction Motors

Discussion and authors' closure of paper 46-116 by C. N. Weygandt and S. Charp, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1000–09.

F. J. Maginniss, N. R. Schultz, and C. Concordia (General Electric Company, Schenectady, N. Y.): This paper is a valuable contribution to our knowledge of the transient performance of induction motors, particularly in the smaller sizes having relatively high resistances as contrasted to the large low-resistance motors treated in reference 4 of the paper.

Expression of the performance equations in terms of the five basic parameters provides for the widest application of the results. The information presented, particularly the speed–response curves, should be very useful to the designer of control and follow-up systems.

The results obtained bring out the importance of low rotor inertia. This is emphasized by the fact that in many cases the load inertia is relatively small so that the motor inertia forms most of the total. Therefore, from the standpoint of machine design we realize that it may be more important to reduce the rotor inertia, even at the expense of a considerable increase in the stator size and in the total machine weight, than to arrive at a minimum over-all motor size.

The authors mention fundamental-frequency speed oscillation in Figure 8a. Actually, these oscillations appear to be about two-thirds fundamental frequency, and as far as we can see there is no reason to expect their frequency to have any simple relation to the fundamental frequency.

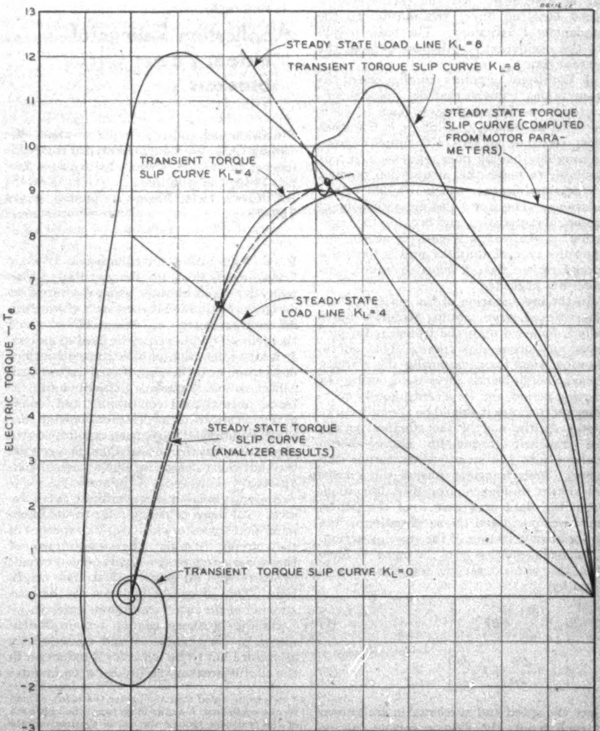
J. W. Momberg (Diehl Manufacturing Company, Finderne, N. J.): In connection with the sixth assumption that the motor is connected to an infinite bus through a zero impedance, it should be noted that this is a very idealized condition. In most servomechanisms one phase of the motor is supplied by an amplifier with rather poor regulation, and the other phase is supplied by the line in series with a capacitor. Only in a contact-making servomechanism supplied from a polyphase source would this assumption be approximately true.

The effects of the rotor time constants plotted in Figure 3 would indicate that rotor inductance should be kept as low as possible. It is for this reason that closed rotor slots seldom are used in squirrel-cage servo motors.

In connection with the conclusion that it would seem desirable not to operate a motor with one phase always energized, it is to be recommended highly that, where it can be arranged, the reference phase not be excited when the motor is not in useful operation. This should be extended to point out that where the reference phase is not energized continuously, the motor can be designed for an even higher ratio of torque to inertia, limited only by saturation and heating effects at the running condition, which is usually much less severe than the stand-by condition with one phase excited. Saturation is rarely the limiting factor in small servo motors, only coming into effect when blast cooling is used. When using the larger servo motors having a maximum output of 200 watts or more, it is almost necessary that both phases be controlled.

It should be pointed out also that the torque of a servo motor rarely increases in direct proportion to the number of poles, because of increased leakage with more poles. In fact, it is not uncommon to obtain the same locked torque for a 4 pole motor as that for a 2 pole motor in the same frame, especially in the small sizes investigated in

Figure 1. Steady-state and transient torque–slip curves computed from the curves shown in Figure 5 of the paper



this paper. Only in large servo motors is it possible even to approach a proportional torque. However, depending upon the application, there may be considerable advantage in increasing the number of poles at the expense of motor performance.

It is hoped that more investigations in the nature of this paper will be forthcoming, since the application of squirrel-cage motors to servomechanisms is relatively new, and there is unquestionably room for improvement.

C. N. Weygandt and S. Charp: The authors wish to thank F. J. Maginniss, N. R. Schultz, C. Concordia, and J. W. Momberg for their discussions on our paper. We would like to reply to the several points brought out individually in these discussions.

Two points in the discussion by F. J. Maginniss, N. R. Schultz, and C. Concordia require clarification. These discussers are quite correct in their statement that the oscillations in speed shown in Figure 8a of the paper are of about two-thirds fundamental frequency. When we first studied this set of curves we had no intention of making the rigid statement that the oscillations are at fundamental frequency. We are in agreement that the oscillations in speed have no direct relationship to the fundamental frequency. The basic reason for this oscillation is unknown to us at the present time.

J. W. Momberg points out that one of the assumptions, namely, that the motor is connected to an infinite bus, is open to some question of validity. While it is true that most servomechanisms are supplied from an amplifier having poor regulation, it was desirable to make this assumption because of the simplification which resulted in our solutions. The study was primarily one of motor characteristics, and because the regulation of the source is entirely dependent upon the type of amplifier and is thus of a special rather than a general nature, this effect was neglected.

In the investigation of the starting characteristics of servo motors, attention naturally is focussed upon the transient torque–speed or torque–slip curve. This curve, showing what happens during the starting period, should be the curve used during the starting period, and for a continuously operating servo motor it should be the only curve used. At the end of the starting period, the transient torque–slip curve should coalesce into the steady state torque–slip curve. Such transient torque–slip curves are shown in Figure 1 of this discussion. From the data of Figure 5 of the paper, there was computed the acceleration of the motor at each instant of the starting period. Then, from equation 36 of the paper we have that the electric torque for this case is given by

$$T_e = 2\tau_m{}^2 \frac{d^2\theta}{dt^2} + 2T_L \tag{1}$$

$$= 2\left[\tau_m{}^2 \frac{d^2\theta}{dt^2} + K_L \frac{d\theta}{dt}\right] \tag{2}$$

Since the speed and acceleration are known at each point, the electric torque can be plotted as a function of time, of speed, or of slip. In Figure 1 of this discussion this torque is plotted as a function of slip. In fact, the torque plotted is the torque actually developed by the motor operating at four times normal voltage. It is seen that when steady state conditions are reached the points representing the electric torques cluster about a region which should be the steady state operating point. The end point of the transient curve should coincide with the intersection of the steady state load line $\left(T_e = K_L \frac{d\theta}{dt}\right)$ and the steady state torque–slip curve. The rather rough nature of the transient curves shown is because of the approximate methods used in the determination of the motor acceleration from the curves of Figure 5. A more precise calculation could be done on the differential analyzer.*

The authors realize that this paper only breaks the surface of the entire servomechanism problem. The next step planned by us is the solution of a simple servo loop, using the induction motor equations where the input voltage to one phase of the induction motor is proportional to an error signal. It is hoped that this second problem will result in a method for representing the induction motor by some source and by an equivalent circuit.

Application Ratings of Indoor Power Circuit Breakers

Discussion and author's closure of paper 46-134 by O. B. Vikoren, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 768–73.

R. C. Van Sickle (Westinghouse Electric Corporation, East Pittsburgh, Pa.): This paper is of considerable general interest to designers of circuit breakers as it gives them information on the conditions under which their circuit breakers may be used in service. It is an explanation of how a rated continuous current can be interpreted in terms of expected maintenance, ambient temperatures, normal load conditions, and emergency overloads. The designer knowing that these conditions vary from application to application has to consider them in a general way but must design for one standard condition.

A circuit breaker is a maximum rated device. All items of rating refer to the maximum performance which can be expected of the circuit breaker. Demonstrations of these characteristics are made on new circuit breakers with all parts in first class condition. This of course includes the demonstration of the rated continuous current.

Circuit breakers cannot remain indefinitely in this condition and consequently allowance has to be made for a reduction in the circuit-breakers ability between maintenance periods and with service. Interrupting ratings are a maximum for the standard duty cycle but occasionally circuit breakers are applied for other duty cycles, shorter intervals between operations, or more operations between maintenance. Tables of per centages for determining the interrupting ratings of oil circuit breakers at various operating duties form American Standard C37.7. O. B. Vikoren applies similar reasoning to the normal current rating and takes advantage of load cycles, low ambient temperatures, and short time overheating of insulation as referred to in the preface to AIEE Standard 1B to get maximum current values at which the circuit breakers can be applied.

In testing circuit breakers for such applications, temperatures of more parts than just the contacts and leads should be taken to be sure that permissible temperatures will not be exceeded. The conductors through insulating bushings are not cooled well and parts heated by hysteresis or eddy current losses may be in contact with oil or other insulating materials.

The conclusion that new apparatus has a better allowance for application beyond its rating than is the case with older circuit breakers, should not be construed as a different basis of design in modern circuit breakers. It is believed that O. B. Vikoren refers to age of the circuit breaker rather than period of design, because the current-carrying ability is influenced less by period of design than by demands of interrupting capacity and speed.

Herman Halperin (Commonwealth Edison Company, Chicago, Ill.): This paper is timely and helpful on the important problem of obtaining the maximum feasible load ratings of equipment.

In Table I of the paper the author indicates a reduction in carrying capacity of 20 to 35 per cent below name plate rating of circuit breakers for a 30 degree centigrade rise of copper-to-copper contacts. This service factor reduction, if any, is obviously a function of the frequency of operation of the circuit breakers. In Chicago the 12- and 4-kv circuit breakers on transformers in attended substations supplying our distribution system are operated practically every other night in connection with shutting down some transformers during the night to reduce core iron losses during light load periods. Such frequent circuit breaker operation means that the contacts are wiped frequently and suffer relatively little oxidation. In some other types of locations the circuit breaker operations are relatively rare and for such cases oxidation will reduce the allowable loading, especially for copper-to-copper contacts.

The author indicates that the total temperature for normal operation should be reduced, for example, from 70 to 60 degrees centigrade when the ambient temperature is 20 degrees centigrade. Although an explanation is given for so reducing temperatures, it is not clear to me. Oxidation is a function of total temperature, and so I think we should learn or estimate what the conditions are for the given circuit breaker involved in each case and then use the same total temperature as the limit for all cases having copper-to-copper contacts.

After equation 3 of the paper the author states that the temperatures rise varies practically as a square of the current. This seems reasonably conservative. We found in testing a few 300-ampere circuit breakers at 300 amperes and at 450 amperes that the final rise with the 450 amperes was close to 2.0 times the rise for 300 amperes, instead of being 2.25 times.

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* It is to be noted that the higher the value of load torque coefficient K_L, the more rapid the approach of the transient torque–slip curve to steady state value. This is in agreement with the curves of Figure 5, which show little or no speed oscillations for high values of K_L.

Under the heading on short-time emergency conditions, the author sets up a total temperature of 90 degrees centigrade as the limit for emergency operation for copper-to-copper contacts. He seems to be concerned about the effect of higher temperatures on the oil and on the annealing of the contact parts. In connection with emergency ratings of transformers, temperatures up to 100 degrees centigrade have been set as a limit for the oil for emergencies. The effect of emergency operation at high temperatures is obviously a function of time. If this operation is to be only for a few hours for each of two days once in several years, then I see no reason from available test data why 100 degrees centigrade should not be a reasonable limit. This is especially true if the circuit breaker is to be inspected shortly after each emergency operation.

This discussion indicates some differences of opinion, and I believe emphasizes the need for obtaining more test data and operating experience at the higher loadings. In the field of underground cable systems, we have obtained in this country enormous amounts of test data and operating experience in regard to the effects of normal and emergency operation at various temperatures. The result is that we have set up temperature limits that are 5 to 45 degrees centigrade higher than used on the other continents. The expense for all the investigations involved has been a very small fraction of the savings obtained. The recent sharp increases in the total installed cost of equipment make it essential to develop a background of information so that starting now we gradually can increase the actual loading of, for example, indoor power circuit breakers.

Philip Sporn (American Gas and Electric Service Corporation, New York, N. Y.): Although circuit breakers generally may not be called upon to carry more load current than their name plate rating, it is apparent from the material in this paper that some circuit breakers in use may not be capable of actually carrying as much load current as the name plate states. As pointed out in the paper, because most circuit breakers are loaded rather lightly this factor has not been serious, in fact it has been so relatively unimportant that it has not obtruded itself upon the average operating engineer's thinking.

The paper offers an approach to problems which arise from time to time with existing circuit breaker installations when loadings exceed those of which the circuit breaker is actually capable, or when it is apparent from studies that loads are going to exceed name plate ratings. The paper also may have application in cases where circuit breakers of sufficiently large current-carrying capacity are not manufactured to meet a given situation. In such cases, actual tests on circuit breakers available can determine what their limitations are; in many cases it may be found they can be applied to meet requirements if judged solely on actual capabilities without regard to ratings.

This naturally brings up another and somewhat more important characteristic of circuit breakers with regard to actualities and their relation to name plate ratings. I have reference to circuit breaker rupturing capacity.

Many circuit breakers in the past were built without adequate design data and they have been sold and used without having a background of actual test. Circuit breakers which have been in use for many years have no actual rupturing test data to show what they will do. Their application hinges on manufacturer's published rating data and oftentimes circuit breakers are discarded or omitted from consideration in specific applications because these data indicate that they are inadequate. This often may not be the case, and considerable sums of money may be saved by making tests on circuit breakers in question to find out what they actually will do. It will be found oftentimes that circuit breakers with or without modifications, are capable of appreciably more than their name plates or published data indicate. They may be capable of much more than the manufacturer's engineers can or are willing to commit themselves.

As one example I can cite a situation at a major generating station where a number of 66-kv circuit breakers of a type no longer manufactured were definitely on the ragged edge as far as rupturing capacity was concerned. Consultation with the manufacturer's engineers indicated that some not strictly orthodox changes could be made in these circuit breakers which might extend their capabilities, but it was not at all apparent to what degree there would be improvement. One of these circuit breakers was rebuilt to incorporate the suggested changes and then was tested on a 66-kv system to the maximum possible extent of that system. Although the manufacturer's engineers would not commit themselves to a capacity of more than 500,000 kva for these circuit breakers, actual test showed that they could handle the maximum that the system was capable of without any signs of distress. This amount was in excess of 900,000 kva.

It is because of experience of that sort that we reached a conclusion a long time ago, a conclusion reaffirmed many times since, that the only sound basis for determining normal and emergency rating with respect to interrupting capacity is by making actual tests. Further, such tests on one type of circuit breaker will not take care of some other type of circuit breaker of the same voltage and rating capacity.

O. B. Vikoren: The author would like to express his appreciation for the interest the discussers have taken in this paper.

Judging from the last paragraph of his discussion, R. C. Van Sickle apparently has read more into the paper than it was intended to cover. Table I of this closure will clarify the statement made under the heading Conclusions of the paper, that is, circuit breakers with copper to copper contacts (generally used on old circuit breakers) have the same temperature rises for currents lower than name plate rating as circuit breakers with silver to silver contacts (generally used on new circuit breakers) have for currents which, at low ambient temperatures, exceed name plate rating.

It should be noted that the paper specifically emphasizes that its purpose is to show the method for arriving at the solution of an application rating problem and that the details may vary for different systems. It is therefore to be expected that Herman Halperin has found conditions different on the Commonwealth Edison Company system as compared with those on the Philadelphia Electric Company system. Nevertheless the temperature limitations should be common for any system particularly when the question arises of establishing a maximum temperature limit for oil circuit breakers. It is believed that Halperin is approaching the danger point by permitting contact temperatures of oil circuit breakers to reach 100 degrees centigrade particularly since oil will begin to smoke at about that temperature. It is believed also that justifying this high temperature on the basis of what is permissible for transformers is overlooking the difference in basic function of the oil in a circuit breaker as compared to a transformer. The arc quenching function of the oil, inherently required in an oil circuit breaker, is referred to here and it is believed that the interrupting ability of a circuit breaker should not be permitted to be impaired for the purpose of stressing the load current limit a few per cent.

Philip Sporn has devoted a considerable portion of his discussion to the matter of interrupting capacity of circuit breakers, which function is important in connection with applying circuit breakers on a temperature basis only so far as the circuit

Table I. Permissible Continuous Load at Different Ambient Temperatures of Two Circuit Breakers of the Same Type and Rating But With Different Contacts

Breakers Installed in Tight Compartments

	Name Plate Rating, 1,200 Amp; Contacts, Copper to Copper; Nominal Rating, 1,200 × 0.8 × 0.9 = 864 Amp			Name Plate Rating, 1,200 Amp; Contacts, Silver to Silver; Nominal Rating, 1,200 × 0.9 = 1,080 Amp		
Ambient Temp, Deg C	Load, Amp	Temp Rise, Deg C	Total Temp, Deg C	Load, Amp	Temp Rise, Deg C	Total Temp, Deg C
40	864	30	70	1,080	30	70
30	935	35	65	1,175	35	65
20	1,000	40	60	1,250	40	60
10	1,050	45	55	1,325	45	55
	$\frac{1,050}{864}$ = 122 per cent			$\frac{1,325}{1,080}$ = 122 per cent		

breaker interrupting ability may be impaired by excess temperatures. It should be noted that the current does not necessarily have to be of shortcircuit magnitude to cause trouble if the opening of the contacts is associated with arcing and the circuit breaker is in an overheated condition.

High Vacuum Leak Testing With the Mass Spectrometer

Discussion and authors' closure of paper 46-142 by W. G. Worcester and E. G. Doughty, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 946–55.

J. E. Burroughs (General Electric Company, New York, N. Y.): Under the heading of "Fields of Leak Detector Application", the paper lists power rectifiers. An interesting example of such an application was the use of the leak detector during the past year on a 3,000-kw 640-volt single-tank multianode rectifier installed in a railroad substation. The leak detector was used to confirm a suspected leak in this rectifier and to test for vacuum tightness after the leak had been repaired.

The leak detector was connected to the rectifier vacuum system by means of a 1/2-inch copper tubing extending from the leak detector inlet connection to a point on the rectifier vacuum system between the mercury diffusion pump and the main vacuum valve. A flanged joint, provided for the rectifier McLeod gauge, was utilized for the leak detector connection. A throttle valve also was installed in the line and was located near the rectifier flange joint.

The leak detector was operated with the throttle valve wide open, because the pressure in the rectifier was less than one micron. This gave an operating pressure measured on the leak detector ion gauge of 1.5×10^{-6} millimeter of mercury to 1.75×10^{-6} millimeter of mercury. Because the residual pressure of the leak detector was 5×10^{-8} millimeter or less, the pressure increase resulting from sample flow was 1×10^{-6} millimeter of mercury or more. A suspected leak from the rectifier water jacket into the rectifier vacuum tank was confirmed by flooding the water jacket with a large amount of helium. The possibility of other leaks on the rectifier was eliminated by hood testing all vacuum seals and welds.

From previous cold seepage tests on the rectifier, the leak rate had been calculated to be approximately 75 micron cubic feet per hour. From the indication on the leak detector, for this amount of leakage it was estimated that the leak detector very easily would have shown a leak of only 1/100 of this value or 0.75 micron cubic feet per hour. The allowable leakage rate for this rectifier tank was 22 micron cubic feet per hour.

After the leak in the rectifier tank had been repaired and the rectifier vacuum system again operating, the leak detector was connected as before, and the water jacket and all vacuum seals flooded with helium. Because the leak detector showed no leaks in the rectifier, it was not necessary to run standard seepage tests as ordinarily done, and the rectifier was prepared immediately for degassing.

From curves which take into account the leak-detector sensitivity, pumping speed of the auxiliary pumping system, and operating pressure, it was indicated that the leak detector would have shown a leak in the final test of the order of 1/10 of a micron cubic foot per hour. This value would be only one-half of one per cent of the allowable leakage rate for this particular rectifier tank.

To summarize the use of the leak detector for leak testing the vacuum system of mercury-arc rectifiers, some of the advantages are listed:

1. No vacuum manifold or auxiliary pumping system is needed with the rectifier because it has its own vacuum system and vacuum pump.

2. The sensitivity of the leak detector is sufficiently great to find leaks impossible with standard test methods, such as cold seepages, pressure, and soap tests.

3. A small leak can be found much faster and located more accurately with leak detector than with the previous standard test methods.

4. The leak testing can be done after the rectifier has been brought up to operating temperature, to find small leaks which do not show up when the rectifier is cold.

5. The leak testing is done with the rectifier under normal vacuum so that conditions are not changed by reversing the direction of pressure difference across the leak, as is done when the tank is put under internal pressure for soap testing.

6. The leak tracing medium, namely, helium, does not contaminate the rectifier vacuum system.

J. G. Trump (Massachusetts Institute of Technology, Cambridge, Mass.): The authors have made an important contribution to the high vacuum engineering art by bringing to an available form the mass spectrometer method of vacuum leak location largely developed by Professor Nier and his associates. This instrument should be valuable not only for large industrial vacuum systems, but also for co-operative use at universities where vacuum is employed in a wide variety of research programs, such as the acceleration of charged particles, vacuum spectroscopy, and electronics generally. Such an instrument would justify itself quickly, both by the saving in time and the superior results in the maintenance of high vacuum conditions. It is to be hoped that the mass spectrometer for vacuum leak testing will be refined still further and reduced in physical size and cost.

Vacuum engineering may be regarded as still in its infancy. Scarcely a decade ago, it was believed generally that pressures much less then 10^{-4} millimeter of mercury could not be produced in large "unoutgassed" metal tanks. In the intervening period, high vacuum production and measurement methods have reached a well-developed stage. Oil and mercury high-speed diffusion pumps, vacuum-tight valves, quick-opening portals, methods of transmitting mechanical power into and out of evacuated chambers, methods of cleaning, conditioning, and "outgassing" in vacuum have become available. Although techniques not fully discussed by the authors, both for leak hunting and for preventing leaks, also have improved considerably, leaks have remained as a serious time-consuming problem, and this new device fills in an almost ideal way the need for a sensitive, discriminating, reliable, and quick-reading locator.

W. G. Worcester and **E. G. Doughty:** The authors wish to thank J. G. Trump for his interesting comments. Widespread interest in the leak detector and its methods of application already has developed. Continued interest and better understanding of application requirements will result in extended usefulness of the instrument.

We also wish to thank J. E. Burroughs for his excellent description of a specific application problem, which is typical of the many successful applications of the instrument.

The authors recognize that other leak-hunting techniques and methods of preventing leaks have been improved; nevertheless the mass spectrometer remains the most sensitive, reliable, and rapid method of leak detection known to us.

There are two general fields of application of the leak detector:

1. Experimental or developmental use to check materials and methods of construction of vacuum apparatus.

2. Production testing of finished products.

Improved techniques in manufacture always can be expected to follow when the exact location and size of leak can be determined readily. Systematic use of the leak detector on production lines as a quality control instrument often will yield savings in time for testing and repairs and improvements in quality of the product.

Control and Protection of Aircraft D-C Power Systems

Discussion and author's closure of paper 46-180 by B. O. Austin, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 741–5.

H. J. Finison (General Electric Company, Schenectady, N. Y.): I am glad to see the emphasis given in B. O. Austin's paper on the provision of and needs for soundly engineered control and protective function for aircraft d-c generator systems. On the whole, I find the objectives outlined by Austin in good agreement with those which R. H. Kaufmann and I have outlined in a series of articles.[1] As Austin has indicated, improvements in 24-volt d-c generator systems urgently are needed to improve performance, reliability, and safety.

Austin indicates that the protective problems are much more severe for 24-volt than for 12-volt systems. I wonder if it is Austin's intent to advocate omitting protection from a 12-volt system or rather to emphasize the greater difficulties that might be experienced with unprotected systems for 24 than for 12 volts. Would Austin please clarify this point?

As Austin has indicated, overvoltage protection is a desirable feature. The magnitude of overvoltage that can be obtained when one generator of a group loses its control of voltage is surprisingly high even with a relatively low strength equalizer. When higher equalizer forces are used (as is sometimes attempted when the generator series winding is used as the equalizer voltage source) the magnitude of overvoltage can be disastrous. Note that with strong equalizer action, the presence of uncontrolled high voltage resulting in high current on one generator tends to force all other generators to a voltage higher than normal. Under many load and speed conditions, failure of one regulator of two or more generators tends to result in votages considerably above 35 and 40 volts and yet without reverse current flowing through the healthy generators. For this condition all overvoltage protectors detect the same overvoltage and destroy selectivity. This factor has made us rather hesitant of using voltage only as a means of selectively determining the unit responsible for the overvoltage. I would appreciate learning whether Austin has experienced the condition I have just described in his development work on the overvoltage protector.

The overvoltage protective device described by Austin is very interesting and appears to possess the desirable element of simplicity. Devices of the type described by Austin are sometimes subject to difficulties with maintaining a constant pickup (minimum trip setting) independent of the position of the plunger. That is, following a transient overvoltage of just insufficient duration to trip the protector, the plunger will have moved almost to the trip position. Because the air gap is now less than at the initial position, it would appear possible that the plunger would be held at the lower position by even normal voltage so that a subsequent moderate transient overvoltage could cause instantaneous trip. Would Austin please describe what methods have been used to overcome this difficulty?

Unquestionably, the combination of balanced current and differential current protection can give a theoretically higher order of sensitivity than usually is possible by use of reverse current protection. Whether this theoretical superiority can be obtained in practice or whether the gain is worth the additional complexity are questionable.

While reverse current protection is incapable of detecting certain types of fault conditions, none of those are likely to occur when the airplane is in flight. And all fault conditions can be detected and corrected by manual supervision. Differential protection does not detect a welded relay while reverse current protection does. In the past, this type of failure was much more prevalent than generator circuit faults. Improvements in generator relay design and in elimination of system disturbances such as those induced by regulator instability, system faults, and heavy motor starting should make the occurrence of welded relays less troublesome, although they still could be as prevalent as circuit faults.

The philosophy of using two devices in series to prevent simultaneous failure usually is predicated on the fact that the failure of one relieves the cause of failure and thereby protects the other. In the application cited by Austin, I wonder if this condition is met. Consider for a moment that a fault exists on the main power system or that a very heavy motor is being started with inadequate capacity. During this condition voltage across the contactor coil is low and current through the contacts is high. If the two contactors are slightly different in characteristics, it is likely that one will weld before the other. But in welding the first contactor does not relieve the condition causing the weld and the continued presence of high current and low voltage may cause the other contactor to weld.

The use of a nonlinear resistor in the equalizer circuit is an interesting development. It should meet a long desired need and permit enjoying high equalizer forces without its attendant disadvantages of excessive influence on system voltage. Would Austin please give us a quantitative figure on the effectiveness of the nonlinear resistor? If the method is effective could not the complication of the additional equalizer relay be avoided with acceptance of the fact that a small decrease in system voltage can be tolerated until manual switching operation disconnects the equalizer of a dead generator?

While the protective functions of differential current, balanced current, and overvoltage relays are fundamental in their effectiveness, I am concerned with the manner in which these devices operate through an auxiliary field relay. This method of operation has the following major objections:

1. Power is required on the system before the generator can supply power to the system.
2. Generator control must be reset every time power is removed from the system.
3. Protective functions require several sequential operations.
4. Protection is relatively slow.

Consider what will happen if a system fault occurs so that the power system is without substantial power for a short period. Sealing power for the field relays of all generators is thereby lost and these relays are free to open. If the system fault is cleared later and if the system load does not pull battery voltage down excessively, the generators can be closed again on the line by manual operation of the reset and control switches. If power stability margins are adequate the generator system may return again to normal operation. Because a manual operation possibly is required before the generators can be returned to supply power, greater battery capacity may be required.

Also consider what will happen if the system fault just described occurs on the battery circuit and in being cleared the system is left without power. Then in spite of the fact that all generators are in good operating condition it would be impossible to make them supply power again because their field relays could not be reset.

Obviously some improvement may be affected by providing sufficient time delay in a field relay with low drop-out voltage to permit it to ride through all likely system faults without opening. This method would have the unfortunate disadvantage of also slowing up the time of the relay for normal operation. In spite of the fact that the nature of the generator circuit permits rapid isolation of a fault by use of superior forms of protection such as reverse current or differential current, in the application portrayed by Figure 6 the delays inherent after sensitive protective devices detect the fault are

1. Field relay time.
2. Contactor opening time.

In summary, I believe that the main protective circuit breaker for the generator circuit should be a device which is latched closed with high contact pressure. Whether the device is closed electrically or manually depends on the weight investment that can be justified for making the circuit breaker remotely reclosable. With such a device it is assured that when needed it will be in operating condition. Devices which are held closed by forces of a solenoid are of questionable reliability when used in a protective function. Further, the protector should be tripped by direct mechanical connection with the protective sensing element. Any attempt to provide the interconnection by electrical means introduces serious problems concerning availability of tripping power.

Reference

1. D-C Power Systems for Aircraft, Parts 1–8, H. J. Finison, R. H. Kaufmann. *General Electric Review*, September 1945 to August 1946.

P. A. Danilson (nonmember; Boeing Aircraft Company, Seattle, Wash.): The need for a positive trouble free protection system is very real. Systems above all must be free from nuisance trips as airplanes have been lost because reverse current circuit breakers tripped out on transients. On multiengined aircraft adequate tripping capacity is very important as fault currents of at least 5,000 amperes can be expected.

Overvoltage Protection

The overvoltage protection system described is satisfactory for an airplane with two generators but is not entirely adequate for one with six generators. Consider one airplane with a maximum continuous load of 1,600 amperes and six 300-ampere generators. Obviously we cannot expect one generator to carry 1,600 amperes and still have overvoltage. However, we can expect a faulty generator to overheat, which in turn will be registered by a thermal indicator light. Protection would operate only as described when system load was near 500 amperes. Because the load on the airplane under discussion very seldom gets that low some other protection scheme appears desirable to prevent the faulty generator from burning up in case the engineer fails to heed the thermal warning light.

Generator Lead Faults and Feeder Protection

The system is positive and fast, the only point in doubt is whether interrupting capacity is sufficient to handle fault currents of a 6-generator system.

Double Break in Generator Bus Feeder

The weight penalty imposed by two contactors should not be necessary. Desirable solution is a contactor that will not stick.

Generator Thermal Indication

Thermal indication is very important on modern transports because generators operate at a high percentage of full capacity whereas on military aircraft the total load was usually less than half the generator capacity.

Control Apparatus

The manufacturer should be complimented on the way he has assembled regulators and relays on a single readily removable panel for easy servicing. Anything that can be done to decrease the number of packages to be installed results in a saving to the aircraft manufacturer. Air transportation companies are demanding that equipment be quickly removable for easy maintenance.

Series Field Versus Ground Resistor

1. The weight advantage is gained because 2.08 volts are available to equalize load instead of 0.5 volt used under Army system.
2. The generators divide the load according to the temperature of the windings.
3. Ballast lamps are required because a larger drop is used.
4. Close production control will be required to assure that the resistance of the series field is held within close limits. If this can be done good load division seems assured.

Conclusion

The system as a whole is a forward step in the direction of improved electrical systems.

B. O. Austin: H. J. Finison and I are in agreement on the needs for soundly engineered control and protective functions for aircraft d-c generator systems. We are likewise in general agreement involving many of my objectives and those of my associates. My associates and I have proceeded actively with the development of a more acceptable d-c power system for aircraft. To date, we have made tests on "mock-ups" which very favorably have verified the contents of the paper.

I have not given any consideration to protective schemes for 12-volt systems. Perhaps there should be more protection than there is in use today. I consider protection against faults in the 24-volt system of prime importance for two main reasons.

1. Twenty-four volts is sufficient potential to maintain low or high current faults involving an arc which readily can become a smoke and fire hazard. Twelve volts will not maintain an arc and therefore is not so much of a hazard.[1]
2. The everincreasing generator capacities involving the 24-volt system further increase the ability of that system to maintain faults high or low at relatively high voltages resulting in exposed electric arcs. This is especially true of low order faults which are below the rating of the generator.

The action of the equalizer ballast lamp in preventing excessive rise in system voltage with full-field operation of one generator is as effective as it is in preventing excessive drop in system voltage in the case of the lowering of the speed of one generator. In fact, under critical conditions the use of the high sensitivity equalizer system with nonlinear ballasts will result in a lower system voltage and a higher reversed current in the second machine than the linear equalizer system using $^1/_2$-volt equalizer voltage at rated generator loads. System tests on an accurate "mock-up" have demonstrated positive selectivity of this system.

An overvoltage relay has not been designed with the characteristics described by Finison. It is an easy task to design the magnetic circuit about the moving plunger and the proper spring characteristics for that plunger, thereby providing a relay suitable for the purpose.

Experiences teach that faults have occurred under all conditions of flight and ground operations. Therefore, protection is needed at all times. Welded contactors in the reverse-current cutout generally are not protected by the balanced current protective system. The reverse-current circuit breaker offers little improvement in this case. In fact, it is a much poorer device when compared with double contactors in the reverse-current cutout. Contactors now used are of the bridge type having two breaks. A long series of tests have shown that this contactor nearly always freezes on one side. When opening under load, the contactor will open a few thousandths of an inch on the free side resulting in a sustained electric arc which burns out that side of the contactor arm. This causes a complete loss of the generator as well as a damaged cutout. The burning of this contactor often happens below the rating of the generator. The reverse-current circuit breaker would not be effected by this condition at all. As previously stated, the double contactor is much more effective, by backing each other up. The frozen contactor usually breaks loose on the reapplication of voltage to its operating coil and continues to operate in a normal manner.

The use of balanced current and differential current protection is a practical, proved arrangement in other fields. It should be applicable and has been proved substantially by extensive "mock-up" tests of recent date. The outstanding feature of this system is the detection of faults below the rating of the generator. It is the low order faults, according to experience, which generate smoke that leaks in through the cockpit and passenger compartments, often causing panic. The low-current high-voltage arcs often last for long periods of time if there is no protection, making fires and smoke a reality.

The reverse-current circuit breaker detects nothing below the rating of the generator. The balanced current system does.

So far as I can see, it is impossible to coordinate properly the opening time of the reverse-current relay contactor and the reverse-current circuit breaker. Timing of the contactors and circuit breakers changes with age and general environment conditions. Balanced current requires no coordination of timing, and environment conditions are not critical with this class of apparatus.

Finison's position supporting manual supervision of fault conditions is not easily tenable, in view of recent airline experience requiring immediate clearing of a fault before smoke occurs.

The philosophy of using two contactors in series is not new in transportation. Finison's example is satisfactory except that his contemplated results apparently are based on contactors and relays with much higher drop-out voltages than those used in the system described. It is only barely possible that either condition described will lower the system voltage below the extremely low dropout of the contactor and adding to that remote possibility that of two contactors in series welding brings us to a relative impossibility.

The quantitative figure of the effectiveness of the nonlinear resistor can be almost anything desired. This depends upon the design of the ballast. The ballast, at the present time, has tungsten characteristics. The equalizer relay can be avoided if one can depend upon the pilot to switch off the dead generator manually. It is not believed that all operators will accept the omission of the equalizer relay.

There are at the present time schematic diagrams and tests which provide a system that can be operated without the battery. This system requires no additional apparatus, but a slight rearrangement in the control circuits. With the use of this system, it is not required that the generator control be reset every time power is removed from the system. Its operation is exactly the same as for systems that are now on the airplanes. The only requirement will be that the pilot will be required to reset when there is a fault.

Actual tests show that the protective functions of the system described are very rapid. For example, a 100-ampere fault is cleared in 0.06 second. The timing of the reverse-current circuit breaker to clear under this condition is infinite.

A greater battery capacity is not required with the use of the current-balanced system. As previously stated, if the operating company so desires, the system can be provided without the use of a battery.

Finison summarizes that the main protective circuit breaker for the main generator circuit should be a latch-closed device with high contact pressures. This, I must admit, is obsolete practice in other forms of transportation and has been so for many years. Double contactors generally have been used instead. I cannot agree that the latched-in type of reverse main circuit breaker always will operate when needed. There are several reasons for this which will be included in the summary here.

Summarizing the discussion, it seems that the following are reasonable facts.

1. The balanced current protective system offers wide range fault protection.
2. The protection against low current fault is of first importance. This type of fault is the smoke and fire maker, as proved by test. High order faults burn clear quicker and are less dangerous, but should be provided for in the protective scheme.
3. Balanced current protection requires no coordination with the reverse-current cutout.
4. Balanced current protection responds only to faults and nothing more, as proved by test.
5. Double contactors for circuit interruption are the modern transportation concept. They lend reliability to continuity of the generator on the system the maximum amount of time.
6. Single contactors have been designed within the last two years with freezing practically eliminated, as proved by test and service.
7. Later schematic diagrams provide for overload protection of the field and control circuits through the balanced current system without added apparatus.
8. Schematic diagrams and tests are in existence providing operations without a battery and reset operations only in the case of a fault. No additional apparatus is required. In other words, this system operates exactly the same as those which have been in use on planes for several years so far as the operator's panel is concerned. The only exception is the reset for faults.
9. The balanced current system is operated remotely and is trip free and can be reset after a fault, provided the fault is cleared.

10. The reliability of the reverse-current circuit breaker is in question. It is subject to changes in calibration resulting from vibration, shock, presence of fine dust particles, and high forward current. The present devices are not trip free.

11. The reverse-current circuit breaker does not sense low order smoke-making faults.

12. It is impossible to co-ordinate the timing of the reverse-current circuit breaker and the reverse-current cutout under several fault conditions. To make matters worse, the timing of contactors and circuit breakers change with age and wear.

REFERENCE

1. GASEOUS CONDUCTORS, THEORY AND ENGINEERING APPLICATION (book), James Dillon Cobine. McGraw-Hill Book Company, New York, N. Y., chapter 4, page 80.

Dielectric Strength of Station and Line Insulation to Switching Surges

Discussion and authors' closure of paper 46-189 by P. L. Bellaschi and L. B. Rademacher, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1047–54.

F. J. Vogel (Illinois Institute of Technology, Chicago, Ill.): This paper should be a material aid in the further standardization of insulation and insulation tests on apparatus, particularly transformers. It often has been quoted that impulse and switching surges were about the same on apparatus, whether it was dry or wet. Would it not have been proper to add to the conclusions that the critical impulse flashover and switching surge flashover under wet conditions might be reduced to 75 or 80 per cent of the dry values?

There is also another question not covered by the paper. The inference is that switching surges are slow and perhaps similated by one-half cycle of 5,500 microseconds duration. Such surges would give nearly uniform distribution of voltage within windings. However, statements have been made several times that resonance could lead to dangerous voltages within windings at the time of and because of switching surges. Since the switching surges were caused by changes in current within the windings, they must be distributed uniformly and the period is long as compared to the transformer period. Agreement on this point, along with the data provided by the author, will help materially in further standardization.

F. O. Wollaston (British Columbia Electric Railway Company, Ltd., Vancouver, Canada): If the wave shown in Figure 5 of the paper were one-half cycle of an a-c wave, the a-c frequency would be about 500 cycles per second. The strength of insulation to a-c frequencies of this order is, as far as I know, about the same as for 60-cycle voltages. But the insulations subjected to the wave of Figure 5 developed impulse strengths. The reason for this could be that the duration of the test wave was essentially only one-half cycle. In that case it could be that insulation subjected to one-half cycle of 60-cycle frequency would develop "impulse" strength (approximately).

I would like to see breakdown tests made at 60-cycle voltage but with the time of application varied from one-half cycle up to the minimum number of cycles that would give breakdown values equal to presently accepted 60-cycle values.

H. L. Rorden (Bonneville Power Administration, Portland, Oreg.): I hope we may not have to take the authors' inference that impulse data are relatively complete to mean that they intend to divert their future efforts to other channels. I believe there is still a vast amount of information to be obtained in this field, and we on the operating end look to the manufacturers for laboratory data. It is significant that we still have insulation failures, and we can only conclude that we still do not fully understand all we need to know about transients. I am sure that the entire electrical profession will agree with me in hoping that Bellaschi and Rademacher will continue their good work. This paper adds useful information to their numerous contributions.

We of the Bonneville Power Administration do not believe that methods of protection have developed to the stage where they are both satisfactory and economical. Although we do not have surge testing equipment in the Northwest, we are conducting such investigations as we find possible without interfering with normal operation of our power system. We differ from most high-voltage systems in the heavy lightning areas of this country by having transmission lines which are not protected by overhead ground wires, although part of our system is in a moderate lightning area. We therefore consider the system itself as a laboratory. The significance of this ties in very closely with the contents of the paper.

The high-voltage lines of the Bonneville system are equipped with overhead ground wires only for approximately one mile from stations and are intended solely for station protection. The transmission lines are susceptible to direct lightning hits, and to our knowledge such hits always result in a flashover. These outages are not of serious concern to us because all of our 230-kv system in the lightning area consists of parallel circuits.

We believe that practically all direct strokes to our lines will result in a flashover regardless of the insulation level. While our normal insulation consists of 16 $5^3/_4$-by-10-inch suspension insulators, we have operated experimentally with as low as 11 on steel towers. Thus far we have found no significant difference in performance. Undoubtedly, enough insulation and conductor separation could be provided to prevent some of the hits from resulting in flashovers, but we believe this insulation level could not be justified economically. We therefore now are considering basing our insulation levels on switching surges. In fact we are planning on an installation involving insulation of this nature and high-speed reclosing breakers in lieu of overhead ground wires. We believe this method of protection may have considerable advantage over the present commonly accepted practice of providing ground wires over the entire system. In such a system we shall have considerable use for switching surge data as well as high overvoltage impulse data. We therefore commend the authors of this paper on the presentation of data which we believe have more than casual importance in the field of transients and insulation levels.

I would appreciate a little additional explanation from the authors on their statement that isolated neutral systems attain five to six times normal voltage. While this is a commonly predicted figure, we have found, all too frequently, that predictions may prove to be too ultraconservative for good practice. While we recognize the fact that isolated neutral systems can be subjected to higher transient potentials than are solidly grounded systems, we are inclined to the opinion that they ordinarily do not attain the magnitude suggested by the authors. From investigations conducted on our 230-kv neutral grounded system, we are confident that our switching surges rarely if ever reach three times normal, and we are contemplating a number of isolated neutral installations for which we must provide adequate protection.

J. T. Lusignan (Ohio Brass Company, Mansfield, Ohio): The authors should be complimented on their contribution to our knowledge of surge conditions on power systems. Their laboratory work has been well done and their presentation of the results clearly shown.

In addition to the insulating members which they already have tested, it would be helpful if they could continue the work as thoroughly into the various forms of limiting gaps used by transmission operators. These gaps range from ordinary rod, horn, and ring gaps to specially designed control gaps.[1] Many of these gaps afford the principal means of protection to station apparatus and the knowledge of their flashover characteristics under wet switching surge conditions should be worth while.

REFERENCE

1. THE CONTROL GAP FOR LIGHTNING PROTECTION, Ralph Higgins, H. L. Rorden. AIEE TRANSACTIONS, volume 55, 1936, September section, pages 1029–34.

H. A. Frey (Locke Insulator Corporation, Baltimore, Md.): The field which has been explored by Bellaschi and Rademacher is one which has been neglected largely by American investigators. The paper constitutes a welcome and timely contribution. As the authors have pointed out, switching surges may be a determining factor in establishing insulation levels for some of the projected higher transmission voltages.

The surges used in the authors' investigations were of about 1,000 and 3,000 microseconds duration. The duration of a 60-cycle half wave is 8,300 microseconds. On such a time scale the effective duration of a $1^1/_2$x40 microsecond impulse wave is almost zero. It is interesting to note that if the curves, Figures 6, 7, 8, and 9 of the paper, be plotted to a linear time scale (abscissa) and if the 60-cycle flashover value be plotted at the time corresponding to the 60-cycle crest, then the switching surge flashover points fall very close to a straight line through the critical impulse flashover and the 60-cycle crest flashover values. In this case the critical impulse flashover is taken as the tail of wave flashover in accordance with American Standards.

Do the authors feel that this could be used safely as a method of estimating switching surge flashover levels from known impulse and 60-cycle flashover values?

J. H. Hagenguth (General Electric Company, Pittsfield, Mass.): The data presented in the paper are a valuable addition to the knowledge of long-duration surge sparkover of insulators and bushings. However, is it correct to call the type of waves used typical switching surges? One type of switching surge shows an initially rapidly-rising front followed by a very gradual rise in voltage with superimposed oscillations. Such waves have more nearly the characteristics of a steep-front flat-top wave in contrast to the slowly rising front wave used by the authors. Before it is possible to draw any definite conclusions as to the sparkover strength of insulators and bushings for switching surges in general, it would seem necessary to have results of tests with the flat-top type of wave. It has been shown[1] that for impulse conditions a large difference in spark-over voltage exists between rectangular waves and straight rising front—as much as 20 per cent at eight microseconds. Such differences in sparkover voltage may not exist at 500 to 1,000 microseconds, the region explored in the paper, but the facts cannot be verified without investigation. For instance, if the test results on insulators or bushings in the paper with spark-over at about 800 microseconds are compared with the straight-line impulse spark-over of similar devices at eight microseconds,[1] the long time spark-over voltage would be reduced to 80 per cent instead of 95 per cent of the short time impulse sparkover.

The fact that the two transformers withstood a large number of short, intermediate, and long duration surges is a proof of the quality of the transformers up to the tested levels, but does not permit a comparison between impulse and switching surge strength of the units since the comparative breakdown values are not known. Further, the damped high frequency oscillations associated with some switching surges when applied to transformer windings which are not shielded adequately are capable of raising internal winding voltages near winding resonance to even higher levels than the impulse.[2,3] The low frequency test wave (Figure 5 of the paper) probably resulted in only straight-line distribution throughout the windings.

The transformer insulation therefore was subjected to less severe stresses than would have resulted from application of flat-top switching surges with superimposed higher frequency oscillations produced externally to the transformer.

REFERENCES

1. VOLT-TIME AREAS OF IMPULSE SPARK-OVER, J. H. Hagenguth. AIEE TRANSACTIONS, volume 60, 1941, August section, pages 803–10.

2. EFFECT OF TRANSIENT VOLTAGE ON POWER TRANSFORMER DESIGN, K. K. Palueff. AIEE TRANSACTIONS, volume 49, 1930, pages 1179–90, Figures 8 and 9.

3. METHODS OF DETERMINING NATURAL FREQUENCIES IN COILS AND WINDINGS, L. V. Bewley, J. H. Hagenguth, F. R. Jackson, Jr. AIEE TRANSACTIONS, volume 60, 1941, pages 1145–50.

P. L. Bellaschi and L. B. Rademacher: The manner in which H. L. Rorden is coping with the insulation protection problem for a system that lies in a moderate lightning area is interesting and indicates a keen realization of the economic aspect of the problem. We fully concur with Rorden that laboratory investigations on impulse and switching surges should continue by all means but in a direction bearing economic importance to the over-all problem. Our experience on switching surges and available knowledge show that for a solidly-grounded neutral system surge overvoltages seldom if ever attain $3^1/_2$ times line to neutral crest voltage, whereas on isolated-neutral systems there has been ample evidence of overvoltages up to five or six times line to neutral crest voltage. For systems of modern design and "good behavior" it is possible that these overvoltages may be even less.

The approach H. A. Frey indicates for estimating switching surge strength from the impulse and 60-cycle values is probably close enough in most cases, but we would hesitate to consider this as a guiding rule. Even for porcelain insulation in air it may be very much in error as evidenced from a consideration of some of the data in reference 12 of the paper, and obviously the method falls far from the mark when applied to oil-immersed insulation.

Tests on other equipment such as J. T. Lusignan has indicated certainly would be worth-while, but we feel that the data already presented provide a good basis from which to assess the significance of switching surge strength in relation to the problem of co-ordination and protection as a whole.

Referring to F. O. Wollaston's discussion the data in Figure 10 were secured in the manner he has indicated. The details of these tests appear in reference 9. In most cases transient voltages are almost over in a matter of milliseconds, and for durations of this order the insulation strength is well up to its impulse level. However, should the system overvoltage approach durations of the order of seconds, then for such a condition the effective strength of the insulation corresponds more nearly to its 60-cycle value.

It might have been well, perhaps, to have added a fifth conclusion such as F. J. Vogel suggests indicating the reduction in impulse and switching surge strength of porcelain insulation in air as affected by rain. However, we feel that conclusion 2 accomplishes the purpose, and that the test data in the paper are specific as to the reduction in strength caused by rain.

Table VI of the paper gives comparative data with reference to a flat-top impulse of the type J. H. Hagenguth refers to in his discussion and should answer the points he has raised in this connection. With reference to the old bugaboo of resonance, we merely want to call attention once again to the ability of modern surgeproof transformers to withstand switching surges of all sorts. This is a matter of extensive experience and record extending over a long period of years.

Those interested in this subject matter both from a standpoint of testing and field experience will find many of the references helpful. The authors have drawn from this over-all experience and their own in planning the investigation and keeping it on a sound and practical basis. In general, we concur with the suggestions and thoughts set forth by the several discussers in regard to the desirability of future studies both in the field and in the laboratory.

An Oscillograph for Recording Transient Recovery Voltages

Discussion and author's closure of paper 46-181 by W. G. Hoover, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1086–91.

E. W. Boehne (General Electric Company, Philadelphia, Pa.): The author is to be congratulated upon his paper which presents an informative and orderly analysis. The recognition and demonstration of the effect of arc voltage and residual arc conductivity upon the recovery characteristics of such high voltage breakers are indeed interesting.

The value of the oscillograms shown in Figures 4 and 6 would be enhanced by a knowledge of the natural frequencies of the circuit and their magnitudes under the more or less ideal conditions of a linear decay of current to zero. A comparison of the phenomena shown in these oscillograms with those presented and discussed in another paper[1] suggests that conduction following current zero is not restricted to the lower voltage breakers.

I would like to ask the author if the test shown in Figure 4C was taken on the same circuit as those shown in Figures 4A and 4B. The 49,000-volt increase about 450 microseconds following current zero suggests a reflection from a junction about 42 miles away. The evidence of such a reflection is not present in Figures 4A and 4B. A circuit diagram would aid in interpreting the films.

REFERENCE

1. THE GEOMETRY OF ARC INTERRUPTION—II, CURRENT-ZERO PHENOMENA, E. W. Boehne. AIEE TRANSACTIONS, volume 63, 1944, pages 384–5, 448.

Laurel J. Lewis (University of Washington, Seattle, Wash.): The development of this oscillograph illustrates in a striking manner the increasing importance to the power engineer of recent advances made in the electronic field. The author is to be congratulated for his use of trigger circuits, pulse shaping circuits, and rectangular pulse generating equipment, more commonly associated with radar techniques, to the development of a new tool for circuit breaker testing.

The writer was privileged to use the instrument described in an extended investigation and can vouch for its reliability. A number of points may be mentioned to emphasize its versatility.

The tests made at the Newark substation of the Pacific Gas and Electric Company, and illustrated by the oscillograms in Figure 4, were among the highest capacity tests made up to that time on an actual system on circuit breakers of that voltage classification. Because of the severe duty imposed on the system it was necessary to limit the number of tests, and only one was made under each test condition. This meant that the oscillograph was required to perform accurately and reliably without any rehearsal. The clearness of the records indicates no small degree of skill and planning in the control of intensity, focusing, and timing of both the vertical and horizontal sweeps.

The problem of accurate tripping becomes of increasing importance as the horizontal sweep speed is increased. Any slight deviation between the time of net flux reversal in the peaking transformer core and the time of the true current zero in the test circuit affects the start of the horizontal sweep. This effect is evident in Figure 4C where the break in the arc voltage on next to the last half cycle appears to occur about 600 microseconds before the position of final recovery. Where the polarity of the recovery voltage cannot be controlled, such a deviation limits the recording space and restricts the speed since proper timing for one polarity may place the final recovery trace on the opposite polarity too far to the right on the screen. Solution of this problem requires that the peaking transformer faithfully follow the current in the test circuit, requiring essentially perfect reproduction by the current transformer of the d-c component in the offset current wave. That such a problem has been solved is illustrated by the oscillograms in Figure 5 where the horizontal sweep speed is about ten times that shown in Figure 4, and the deviation has been reduced greatly.

The applications described are those in which one terminal of the circuit breaker is grounded, and it is thus possible to keep the recording equipment essentially at ground potential. This requires that the recovery voltage be applied to only one plate of the cathode ray tube with the other maintained near ground potential. Ordinarily this contributes to a certain defocusing effect in the tube. Perhaps Hoover could elaborate on how some of the difficulties from this cause have been overcome. It would also be interesting to know if the oscillograph can be used on line-to-line tests.

W. G. Hoover: In reply to E. W. Boehne, the natural frequencies corresponding to the tests of Figure 4 were not known nor were they calculated prior to the tests. A rough estimate of their order of magnitude was made to determine the necessary frequency response characteristics of the oscillograph circuits. In the tests of Figure 6, the high frequency oscillation in Figure 6A was mainly caused by a capacitance of about 370 micromicrofarads (breaker, potential divider, leads, and reactor distributed capacitance) and a series reactor of about 1.3 millihenries. The generator side of the reactor was connected to a 1,050-kva sine-wave generator through about 100 feet of single-conductor 15-kv lead-covered cable. The low frequency oscillation observable in both oscillograms is caused principally by the generator reactance and equivalent total terminal capacitance.

The test shown in Figure 4C was taken on the same 110-kv bus as those of Figures 4A and 4B. In the case of Figure 4C only, the three conductors of an otherwise idle 110-kv transmission line were connected to the high voltage terminal of the test breaker for the purpose of reducing the transient recovery rate. The reflection from the open end of the line about 40 miles away has no counterpart in the tests of Figures 4A and 4B.

The author appreciates the comments made by Laurel J. Lewis who was present at and contributed considerably to the success of the tests described in the paper. The question of beam focus for deflections over the whole face of the cathode-ray tube was treated satisfactorily by using push-pull deflection voltages for both vertical and horizontal sweeps. The signal voltages were impressed on only one of the vertical deflection plates for line-to-ground tests, but the amplitudes of these voltages were maintained low enough so that serious defocusing did not result. Tests made during the early development of the oscillograph showed the complete feasibility of measurements on line-to-line tests, wherein the outputs of identical voltage dividers connected to each terminal of the test breaker are brought to the separate vertical deflection plates.

Long-Distance Power Transmission as Influenced by Excitation Systems

Discussion and authors' closure of paper 46-182 by C. Concordia, S. B. Crary, and F. J. Maginniss, presented at the AIEE Pacific Coast meeting, Seattle, Wash. August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 974–87.

R. D. Evans: (Westinghouse Electric Corporation, East Pittsburgh, Pa.): In their present paper Concordia, Crary, and Maginniss deal with two subjects, namely,

1. The relation of exciters and regulators to the stability of transmission systems of the moderate length of 200 miles.
2. The relation of intermediate synchronous condenser stations to the steady-state stability of long transmission systems of 600 miles in length.

These subjects have received relatively little attention during the last few years but they were discussed actively during the years from 1924 to 1929. It is interesting and sometimes valuable to review basic problems at intervals of approximately 20 years, as such intervals provide time for the development of new methods of analysis, reconsideration of basic assumptions, and the production of new apparatus and modes of operation.

The first subject considered by the authors is the effect of the characteristics of exciters and regulators upon the stability limits of straightaway transmission systems for both steady-state and transient conditions. The results of this part of the investigation by the authors are summarized in the paper under the heading "Conclusions," items 1 to 4. Item 3 seems to state that the transient-stability limits for conditions imposed by line faults with subsequent high-speed sectionalizing are lower than those for the steady-state conditions after the switching operations. Item 1 states that for a system subjected to a severe fault with the subsequent switching operations, the stability limits cannot be increased appreciably by practicable changes in the exciter and regulator system. These conclusions are not new; they have been held since high-speed fault-clearing means became a reality. Item 2 seems to state that for certain less severe faults or for faults of shorter durations, some relatively small increases in stability are possible by the use of excitation systems of higher response. This conclusion, while interesting, appears to be of academic value in view of Item 1 which states that little benefits are obtainable for the more severe conditions which are controlling. Item 4 states that with a properly stabilized excitation system and with stability maintained on the first swing, stability generally should be maintained on the subsequent swings. In this connection it is pertinent to observe that practical systems will be laid out with a margin between the load to be carried and the stability limit. When systems are operated with such margins, the chance of instability occurring on subsequent swings will be minimized. If the stability studies had taken the normal load margins into account, the results would have been of greater practical interest.

The second subject considered by the authors deals with the increases in the steady-state stability limits that result from the use of regulated synchronous condensers located at intermediate points on a long transmission system. It is of interest to compare the results given in the present paper with those presented by C. F. Wagner and R. D. Evans in an earlier paper (reference 7 of the paper). The layouts assumed for the stability studies are different but it is of interest to compare the per cent increases in the stability limits for one and for two intermediate condenser stations. This may be done by taking the data from Figure 14 of the paper and replotting to the form used in Figure 26 of the earlier paper by Wagner and Evans. It will be found that quite good agreement will obtain for both one and two intermediate condenser stations of low reactance. For increasing values of condenser station reactance, the checks are not as good. This is a result to be expected from the differences in the layouts assumed for the studies.

The hypothetical system studied by Concordia, Crary, and Maginniss includes finite reactances for the hydroelectric generating station, the sectionalized double-circuit 230-kv lines, and a receiver system represented by a synchronous motor. The hypothetical system assumed by Wagner and Evans is similar except that the receiver system is represented by an infinite bus. The practical system probably would involve a step-down station, large shunt loads, and receiver-end generators. On the practical system the receiver-end generators will have a phase leading that of the shunt load. It would appear that the approximation of an infinite receiver is not only simpler but quite as accurate as one using a synchronous motor without a shunt load. However, either approximation can be used, and both will show that important increases in the stability limits are possible through the use of intermediate synchronous condenser stations. In any actual problem these differences in the approximations would not appear because the actual system would be represented accurately.

Alexander Dovjikov (Bonneville Power Administration, Portland, Oreg.): The proposed increase in the range of transmission of power supplied by the Bonneville Power Administration has made it necessary to reconsider ways and means of accomplishing this purpose in the most reliable and economical way. In parallel with considerations of d-c transmission, of increase in a-c operating voltage, and of the series capacitor-compensated lines, interest in intermediate synchronous condensers and in the importance of interconnections with other sources of electromotive force along the long line such as local generating stations, has been revived. This revival is noticeable not only in this country but abroad as well.[1,2] The paper by Concordia, Crary, and Maginniss is therefore very timely, and is of great interest to the engineers of the Bonneville Power Administration.

Beside the preponderant importance of the speediest fault clearing for increasing the transient stability of a transmission system, the value of appropriate intermediate synchronous condensers (known as the Baum system) in increasing steady-state and transient stability of long distance transmission systems has been realized for at least the last 25 years. The stabilizing effect of such synchronous condensers is a function of their ability to maintain their terminal voltage, and it was realized early again that to achieve this the machine must have:

1. Low reactance.
2. Very high-speed excitation system.

Or in other words, the synchronous condenser must have voltage characteristics analogous to a d-c machine with compound excitation. Excitation of such an a-c machine (in addition to the no-load excitation) must be a function of load current and of load power factor, and, accordingly, the late Doctor C. L. Fortescue proposed the use of:

1. A cylindrical rotor with a 3-phase distributed field winding, the terminals of which through slip rings would be supplied by three direct currents from

2. A 3-brush rotary convertor rectifying 3-phase alternating currents from the secondary windings of current transformers, the primary windings of which are inserted at the terminals of the synchronous machine. The field produced by these three direct currents is proportional to the load, and the direction of this field depends upon the load power factor.

By changing the ratio of the current transformers this machine may be made over compounded. The excitation response is instantaneous and the machine may keep its terminal voltage constant in the limits of its rated capacity.

The construction of this machine proved to be too expensive. As the cost of the conventional excitation system is a very small fraction of the cost of synchronous condensers, and any additional cost in connection with its improvement will be reflected only slightly in the cost of the total installation, efforts of designers again were switched to the development of very high-speed excitation. Several measures, including lamination of stators of d-c exciters and of special quick acting contacts on automatic voltage regulators, produced quick response excitation.

In the selection of synchronous condensers located in the main 230-kv loop of the Bonneville Power Administration transmission system, we recognized that in addition to their voltage regulating activity on the substation bus and the control of reactive power flow through the points of interconnection with other system, these machines are intermediate condensers. Together with the attempts to obtain machines of low reactance, we were increasing consistantly the kilowatt capacity of exciters and particularly their ceiling voltage. Our latest specifications for 35,000-kva synchronous condensers required:

> The main exciter shall be a 250-volt direct-current generator with a capacity at least 20 per cent greater than that required for the excitation of the condenser when it is carrying full kilovolt-ampere load. The ceiling voltage of the main exciter, when delivering full rated current, shall be not less than 175 per cent of normal rated voltage. The main exciter shall be designed to have a nominal exciter response of not less than 1.5, as defined and measured in accordance with American Standard *A*50-1936, paragraphs 1.219, 1.220, and 1.221. The pilot exciter shall be a 125- or 250-volt direct-current generator with a capacity at least 50 per cent greater than that required for supplying the field current of the main exciter, when the main exciter is carrying rated current and voltage.

If the system of excitation using an amplidyne is not much more expensive than a conventional d-c exciter, we believe it may be used very profitably. The curves presented in Figure 9 of the paper, however, did not impress us so much with the effectiveness of the high speed of excitation as by the necessity of shortening the time of fault clearing. Our trouble is—and we believe it is shared by other operating companies—that although we installed the fastest relay systems (distance relays with carrier control) available, known as instantaneous, the time of fault clearing on both ends is in the region where Figure 9 shows the use of very high-speed excitation to be practically useless. We are not as pessimistic in this respect as the authors of the paper, although we agree with them in general; however, we believe that relay systems must be developed which will be very fast. Could the authors make any statement regarding the progress being made in this respect? There is another statement in the paper in which we were interested even more than in the high-speed excitation, namely, in the necessity for decreasing the transient reactance of the synchronous condensers. No excitation could be so fast as to neutralize the instantaneous terminal voltage drop resulting from the transient reactance, and we fully agree with the idea which was advanced very clearly 20 years ago by Evans and Wagner[3]

> . . . machine's ability to supply lagging kilovolt-amperes with drop in voltage is a measure of its desirability from the standpoint of maintaining stability. To supply this large amount of lagging kilovolt-amperes with drop in voltage, a synchronous machine should have low leakage reactance, as this is initially the limiting factor, and in addition should have a "strong field and weak armature" which tends to prevent reduction in machine flux when supplying the large amount of lagging kilovolt-amperes. Merely derating a synchronous machine will not give as favorable a result in cost and performance as can be obtained by specially designing the machine. Similar measures can be applied to synchronous generators with corresponding improvements in performance.

Strangely enough, machines with such characteristics have not been designed as yet, and our consistent efforts to obtain low reactance synchronous condensers always have been countered by their high price. Evidently, factory cost estimators were "merely derating a synchronous machine"—procedures against which their own general engineers were objecting. We were glad, therefore, to see the following statement in the present paper:

> The building of low transient reactance synchronous condensers with special excitation systems now appears entirely practical, and makes it necessary to consider such a method in the design of any new long-distance transmission system. However, if it were not possible to attain the special features, the required synchronous condenser capacity becomes extremely large, making this method uneconomical compared with other methods.

This infers that these special low reactance machines can be produced at costs approaching those of conventional machines.

What we would like to ask the authors, who represent the largest organization for manufacturing electric machinery is whether such synchronous condensers were built or at least designed, in what respects this special design would differ from standard, and how the cost of such a special low reactance machine would compare with the costs of standard low reactance machines and with machines of the same kilovolt-ampere rating having the usual high reactance.

REFERENCES

1. Long-Distance Transmission of Energy and the Artificial Stabilization of Alternating-Current Systems, G. Darrieus. Confèrence Internationale des Grands Réseaux Électrique à Haute Tension (Paris, France), 1946, paper 110.

2. Alternator Exciter Systems, R. David. Conférence Internationale des Grands Réseaux Électrique à Haute Tension (Paris, France), 1946, paper 101.

3. Static Stability Limits and the Intermediate Condenser Station, C. F. Wagner, R. D. Evans. AIEE Transactions, volume 47, 1928, pages 94–121.

C. Concordia, S. B. Crary, and F. J. Maginniss: R. D. Evans' re-emphasis of the previous work done many years ago and wherein it checks with the present work is appreciated very much. Our objective was to present sufficient additional quantitative information in terms of present-day apparatus so as to lend encouragement to the reconsideration of the use of intermediate synchronous condensers as a practical means for increasing the power limits of long-distance transmission systems.

Alexander Dovjikov raises a question in regard to instantaneous relaying. Reliable relays of not more than one cycle total operating time are now generally available for high-voltage transmission systems. Higher speeds of relay operation would be of small stability benefit and would be of questionable value if the high speeds were attained at a sacrifice of relaying reliability.

Dovjikov also asks about the use of special synchronous condensers. It is difficult to give any cost comparisons because the cost of such condensers depends not only on the required reactance but also on the required capacity in a single unit. Cost comparisons which we have made for a long high-voltage line requiring intermediate synchronous condensers having an effective reactance of approximately 15 per cent on a 400,000-kva base indicate that the cost would be entirely competitive with other methods of line compensation in order to obtain the same steady-state stability limits. We would like to encourage Dovjikov to make a similar study and to make his comparisons for a specific case for which cost estimates may be obtained.

Field Tests on Power-Line Carrier-Current Equipment

Discussion and author's closure of paper 46-199 by R. H. Miller and E. S. Prud'homme, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, December section, pages 824–7.

F. S. Benson (Pacific Gas and Electric Company, San Francisco, Calif.): Since this paper was prepared we have installed between Shasta substation and Pit 5 generating station an additional carrier current telephone channel using a frequency-modulation carrier installation. This later installation has proved quite satisfactory. Our operators and dispatchers state that the background noise level has been reduced materially, and that the fidelity of tone is good, making telephone conversations much easier to understand. Peculiarly enough, there seems to be a greater reduction in background noise level at the Pit 5 end than at the Shasta substation. At the moment, we do not know how to account for this. This carrier current installation operates over a rather high-loss 220-kv circuit; the corona loss is high because of the type of rope core conductor originally installed. Our frequency-modulation carrier installations are giving good service now, particularly since the adjustments mentioned in the paper have been made, and have solved a troublesome background noise problem.

S. C. Bartlett (American Gas and Electric Service Corporation, New York, N. Y.): The paper presents information on a subject on which all too little specific technical and experimental data are now available. In radio applications the relative advantages of frequency modulation and amplitude modulation have been investigated rather thoroughly and theoretical studies and field observations well correlated and generally accepted throughout the industry. The application of frequency modulation for carrier current communication on a practical power system has introduced a number of conflicting viewpoints. Information of the type presented in this paper is of value in checking the purely theoretical reasoning. However, there are a number of questions which must be answered before we can employ intelligently any new technique by having complete information regarding its advantages and limitations. Additional data along the lines of that presented in this paper would help considerably.

The authors have produced a very interesting and enlightening paper but there are several points not exactly clear. For instance, if we interpret properly Table II in which frequency modulation and amplitude modulation signal-to-noise ratios are compared, it appears that as the signal-to-noise ratio diminishes (reducing transmitter power), the advantage of frequency modulation over amplitude modulation also decreases. This, of course, is to be expected, because if this process were carried to the point where signal and noise were equal there would be no advantage of one system over the other. The important point is that at low signal-to-noise ratios, where amplitude modulation becomes unsatisfactory, frequency modulation still provides a means of improvement and might transform an unsatisfactory channel to a serviceable medium. It would aid considerably in interpreting the results if the authors would indicate what from their experience constitutes the minimum signal-to-noise ratio at which intelligible communication can be obtained. Also in order for a frequency modulation receiver to perform properly, it is essential that the limiter be saturated. We wonder if the receiver used for the tests had sufficient sensitivity so that this condition was realized for the weakest signal recorded.

In this matter of deviation ratio it has been assumed generally that for our purposes the deviation must be limited to one-to-one. This is doubtless correct for main transmission systems where the carrier spectrum is crowded with channels of all types. However in a great many cases such as subordinate transmission systems requiring only one communication channel and possibly a few relay channels, it would be entirely feasible to increase the deviation ratio if there would be a corresponding advantage. For radio work we know this would be true, but on a power line the transmission characteristics are such that the sharp peaks and valleys might cause serious distortion of a wide-band system. The authors discuss only results of tests using ratios up to one-to-one, but if observations beyond here were made such data should be included.

In connection with existing amplitude modulation channels where signals are of adequate magnitude, and considering the difficulties of converting to frequency modulation, it is reasonable to believe that amplitude modulation will continue to be used indefinitely. For this reason the authors' comments regarding amplitude modulation receivers are quite interesting. Specifically, it was mentioned that fidelity sometimes was impaired by the high Q of the line tuners and could be improved by broadening the tuning. This, of course, can be done in many different ways, and we believe it would be interesting for the authors to outline the method used and possibly the reasons for the choice of method. Evidently the receiver used had extremely high selectivity to start with, because it has been our experience that it is frequently necessary to sacrifice some audio fidelity to obtain the necessary selectivity. In many cases we have made it a practice to take advantage of the high Q tank circuit of the transmitter to provide additional receiver selectivity by tapping the receiver to the tank circuit rather than to the radio-frequency line. This, of course, applies only to single frequency sets.

Mention is made that alignment of amplitude modulation receivers should not be made by adjusting for peak detector current. It would be valuable to have the authors' opinion as to the manner in which such tuning should be accomplished.

The electric utility industry needs specific answers to two fundamental questions in connection with the use of frequency modulation carrier. These are

1. **Where an existing amplitude modulation system shows only marginal performance because of noise, will conversion to frequency modulation using the same frequency, power, and a one-to-one deviation ratio convert this to a serviceable channel?**
2. **In planning new channels where it is reasonably certain that amplitude modulation will provide adequate signals, does frequency modulation offer advantages enough to offset the additional cost and complication?**

Any pertinent information along these lines would aid materially in planning future additions or modifications.

In general, it is our feeling that the paper is a valuable contribution to our limited knowledge of comparative frequency modulation and amplitude modulation performance on a practical power system. The only general comments we have is that the paper, if anything, is too lacking in details and we hope the authors as well as others will be encouraged to make available whatever additional information is obtained along these lines.

R. C. Cheek (Westinghouse Electric Corporation, East Pittsburgh, Pa.): The authors have drawn two conclusions that apparently are not supported by the data presented in the paper. Item 3 under "Conclusions," based on the data presented in Table I of the paper, states that "frequency modulation equipment is capable of maintaining a high signal-to-noise ratio with large variations of carrier signal." The data given in the table indicate that the signal-to-noise ratio was quite high for all tests, so that the figures given in the second column, received volts, can be used as a measure of relative received signal strengths, even though a small amount of noise is included. An increase in the received signal level from 1.5 volts to 2.2 volts would be expected to raise the signal-to-noise ratio by 3.3 decibels for any system, and this is precisely the change in signal-to-noise ratio obtained by the authors for this increase. One would expect the signal-to-noise ratio to increase by about 4 decibels with a change in the received signal level from 2.2 to 3.6 volts. The authors obtained an increase of 1.7 decibels for this change, giving them a signal-to-noise ratio of 29.5 decibels instead of 32 decibels as would have been expected. This discrepancy, however, may have been a result of slight changes in the actual noise level during the test.

It therefore appears that the frequency-modulation system signal-to-noise ratio is no more independent of received signal strengths than other systems. In addition, the noise reducing properties of the frequency-modulation system depend upon the maintenance of a received signal voltage considerably in excess of the noise voltage, in order for the limiter to be effective. Even with perfect limiter action, however, the equivalent frequency-modulation resulting from phase modulation of the received carrier by noise voltages is directly proportional to the relative strengths of noise and received signal. Thus in frequency-modulation as in any other system, a change in the received signal level of so many decibels will give a change in signal-to-noise ratio of the same number of decibels, provided the frequency modulation signal is always several times greater than the noise. If the latter condition is not met, a slight reduction in the frequency modulation signal strength may reduce the signal-to-noise ratio by a much greater amount. A comparison of the relative strengths of transmitted signals with received signals on successive tests in Table I of the paper indicates considerable nonlinearity in the carrier channel. It would be interesting to know what explanation the authors have of this phenomenon.

The authors state in item 4 under "Conclusions" in the paper that the frequency modulation system "permits successful carrier current operation on lines with noise levels so high as to preclude the use of conventional amplitude modulated equipment" basing this conclusion on Table II of the paper. It is unfortunate that the tests shown in the table were not made for smaller signal strengths than those used, because quite the opposite result no doubt would have been demonstrated. As a matter of fact, even though the test was carried only to the point where the signal voltage was still at least 7.5 times the noise voltage, a very definite reduction in the gain of frequency modulation over amplitude modulation was evident. In other words, the frequency modulation system provides gains over amplitude modulation under favorable noise conditions, where the noise voltage is already small compared to the signal, but provides less gain and may even give a loss in signal-to-noise ratio when compared with amplitude modulation or single side band under severe noise conditions.

Although the authors did not state in the paper that the frequency modulation equipment used in the tests included audio pre-emphasis and de-emphasis circuits, while the amplitude modulation equipment did not, it is assumed that this was the case. The figures given for the gain in signal-to-noise ratio of frequency modulation over amplitude modulation therefore must be reduced by an amount equal to the gain obtained by pre-emphasis, if a true comparison of the two fundamental systems is to be made. It should be remembered that pre-emphasis can be used with an amplitude modulation or single side-band system, and that, although a given amount of pre-emphasis provides a greater reduction in noise with frequency modulation, the amount of pre-emphasis that can be used with a frequency modulation carrier system is limited because of the increased band width required by the signal as higher voice frequencies are emphasized to a level approaching that of the lower voice frequencies. Such an increase in band width does not occur in an amplitude modulation or a single side-band system, so that more pre-emphasis can be used, giving gains from pre-emphasis comparable to those obtainable with frequency modulation.

The effect of high Q line tuning equipment upon the audio-frequency range of a carrier channel is worthy of additional comment. The reduction of higher order side bands by such sharply tuned circuits will reduce the response of an amplitude modulation channel to the higher audio frequencies, as described by the authors. The corresponding effect in a frequency modulation channel is not primarily an alteration of the frequency response curve, but is the introduction of harmonic distortion at all audio frequencies. The apparent remedy used by the authors was to broaden the carrier-frequency response curve of the tuning equipment, with a corresponding reduction in signal strength at the carrier frequency, and an increase in the noise voltage accepted by the receiving equipment. It is interesting to note that a single side-band channel, because of the extremely narrow band width it requires, permits the use of tuned circuits of higher Q than can be tolerated with either a frequency modulation or an amplitude modulation channel and consequently does not require such compromises in signal-to-noise ratio for the sake of increased band width.

E. S. Prud'homme: R. C. Cheek's comments hinge largely around some important and rather drastic misinterpretations of the data in the paper. It appears desirable to hasten correction of these because the data, properly explained, seem to support adequately all of the conclusions.

Referring to the second column of Table I of the paper, it was pointed out that the received volts are signal *plus* noise. It is not proper to use these values as a measure of relative received signal strengths. A curve of received volts against transmitted volts will show clearly that approximately 1.4 volts of noise did exist with no transmitted voltage. This was not stated in the paper because it was thought this should be apparent from the figures given. Because the signal and noise voltages do not add arithmetically on a vacuum tube voltmeter, the curve will not be linear. Therefore, there is no reason to attempt an explanation of any phenomenon of nonlinearity. The signal-to-noise in decibels figures were measured, of course, at the receiver output terminals, and these values are not related directly to the received volts data which were a total voltage measured at the input terminals, and included both signal and noise. The calculations of Cheek, prepared upon the false assumption that these figures can be used as a measure of relative received signal strength, therefore, are misleading.

The conclusion that frequency modulation permits successful operation with noise levels so high as to preclude the use of conventional amplitude modulation equipment, was arrived at in a very practical manner—acceptable operation was not obtainable with 10 watts of amplitude modulation and was obtained using a 2-watt frequency modulation unit. When it is understood that a high signal-to-noise level is desired, it is evident from Table II of the paper that frequency modulation equipment gave a more acceptable ratio with an output of 0.2 watt than did the full 10 watts of amplitude modulation.

The authors restricted themselves in this brief description of the field tests to the portion of the data that would be of wide interest and of practical value. The variation in the frequency modulation carrier signal was more than 100 to 1, which the authors considered to be large (the frequency modulation equipment had a rated output of 25 watts; the comparison terminated at 10 watts which was the maximum rating of the available amplitude modulation equipment).

From the preceding explanation and the first paragraph of the paper, it should be evident that this section of line presented very unfavorable noise conditions, markedly so as compared with several other line sections on this same system. The representative data included in the paper cover smaller signal strengths than were of any practical operating value even for the frequency modulation equipment. As a matter of fact, some of the unreported tests extended to signal strengths at which the amplitude modulation signal was completely unintelligible, while the frequency modulation signal remained intelligible, and none of these tests indicated to us the opposite of any of our conclusions.

The frequency modulation equipment did use pre-emphasis, and the amplitude modulation equipment did not. There was available for these tests neither amplitude modulation with pre-emphasis nor frequency modulation without it. It would be interesting to see comparable field data taken with amplitude modultion equipment using pre-emphasis.

The loss in signal-to-noise ratio resulting from the lowering of the Q of the receiver line tuning equipment was so small that no additional discussion of this point was included in the paper. While it is true that flattening the carrier frequency response curve of the receiver line tuning equipment does reduce the magnitude of the received carrier signals, it also reduces the magnitude of noise voltages at the same frequencies. Only the noise voltages near the carrier frequency are important as the tuned circuits in the receiver will reject noise signals of other frequencies. Therefore, the reduction in received carrier signal does not result in a significant loss in signal-to-noise ratio.

Characteristics of a 400-Mile 230-Kv Series-Capacitor-Compensated Transmission System

Discussion and author's closure of paper 46-196 by B. V. Hoard, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1102–14.

R. D. Evans (Westinghouse Electric Corporation, East Pittsburgh, Pa.): This is an interesting and valuable addition to the art of long-distance transmission. It should be considered in connection with an article by S. E. Schultz.[1] Hoard in his paper deals with the technical problem of long-distance a-c transmission, particularly in regard to system stability, use of series-capacitor compensation, and line relaying. Schultz in his article considers the more general aspects of the problem as to the choice of system and economic considerations. He concludes that the 230-kv a-c system with series-capacitor compensation is one of the most promising means of obtaining low-cost long-distance transmission, and that its further investigation should be continued.

I wish to comment briefly upon the choice of system for the 400-mile transmission problem considered by Hoard. In the present state of the art it is my opinion that the a-c system is to be preferred to the d-c system because

1. The a-c system is undoubtedly more suitable for supplying a network or for supplying loads at intermediate points along the line.
2. The prospective savings with d-c transmission appear to be small.
3. The adequacy and cost of the d-c system can be established only by a large-scaled installation using rectifying apparatus, the development of which is incomplete.

The selection of the transmission voltage depends upon many factors, including the prospect of a transmission network, the availability of wood poles in the area under consideration, and the technical considerations affecting the dimensions of steel towers, line conductors, and ground wires. The choice of 230-kv transmission in preference to higher voltage may not be as advantageous in other localities, particularly when steel towers are required.

In Hoard's paper the stability limitations inherent in long-distance a-c transmission are avoided by the use of series-capacitor compensation. Series capacitors used are of the protected type with means for quickly restoring the capacitor to the circuit after fault clearing. This method of using series capacitors was described first by E. C. Starr and R. D. Evans[2] (see reference 1 of the paper). While previous papers proposed the application of series capacitors to lines of various lengths, Hoard's paper is the first devoted to a system with a 400-mile line. This system has been studied extensively with the aid of the a-c calculating board as shown by the results of stability studies reported in the paper. For a 400-mile transmission system with series capacitors, the use of two groups appears advantageous. The line-fault clearing times of $3^{1}/_{2}$ and 7 cycles appear reasonable and practical. The total time of $10^{1}/_{2}$ cycles assumed for the restoration of the series capacitor to the circuit appears commercially feasible.

One aspect of the study reported by Hoard will require further consideration, namely, the line-relaying problem on a transmission system with series capacitors. The use of the current-balance scheme seems to have the advantages ascribed to it, but it may be practicable to use other schemes. In this connection, the detailed data in the paper as to the operating voltages and currents at relay locations should be of great value.

The assumptions made by Hoard for the stability and relaying studies on the 400-mile transmission system undoubtedly were based on prospective developments of the Bonneville system. However, these assumptions are widely applicable. More specifically, the assumed layout takes into account the impedance of a hydroelectric generating station, a sectionalized double-circuit transmission line with two intermediate series capacitors, and a receiver step-down station with shunt load and receiver-end generators. This type of layout is more generally applicable than the one sometimes used in which the receiver is represented by a synchronous motor. Moreover, the type of fault assumed is that most generally accepted as the criterion of system design from the stability standpoint, namely, the double-line-to-ground fault. These circumstances increase the usefulness of the results for application to other systems.

Reference

1. 500-Mile 230-Kv Series-Compensated Transmission, S. B. Schultz. *Electrical World*, New York, N. Y., August 17, 1946, pages 70–1.

S. B. Crary (General Electric Company, Schenectady, N. Y.): Although previous papers have indicated the general possibilities and field of application of series capacitors for long transmission lines, Hoard's paper properly and clearly points out some of the important problems involved in protection, control, and line relaying. This is a natural result of getting down to a specific case. The author is to be commended for presenting this type of analysis.

Somewhat different conclusions as to the relative importance of the various factors may have been obtained if the conclusions had been based more on simultaneous fault clearing than on sequential fault clearing. Simultaneous fault clearing is even more important for the type of line studied than for conventional uncompensated lines and therefore should be considered a necessary requirement. It was natural, however, for the author to make his study based on equipment which generally is known to be available such as balanced current protection. If simultaneous fault clearing is used, the following results may have been obtained as contrasted with those results of the author based on sequential fault clearing:

1. Lower-than-normal generator reactance and higher-than-normal generator inertia should be used to reduce the required line compensation for a given line loading. Lower reactance becomes more important than higher inertia with the reduction in total fault clearing time obtainable with simultaneous clearing.
2. For a given line loading and distance, the compensation required becomes less with simultaneous fault clearing. A further reduction in the required capacitor kilovolt-amperes and line compensation can be expected also by the use of low-transient reactance generators.

For these reasons, the author's total required amount of compensation as well as the percentage increase required to allow for nonremoval of the capacitor during the transient may be actually somewhat larger than necessary. A further reduction in allowable compensation is in the direction to simplify and make more practical the use of series capacitors.

In conclusion, simultaneous fault clearing rather than sequential fault clearing leads to such important advantages and tends to change the basic relation between the important system design factors to such an extent that it must be considered one of the more important requirements of a series-compensated long-distance transmission system.

For a single-circuit line rather than a multi-circuit line, the relaying and protection difficulties for series compensated lines largely disappear. This type of installation may find application for other cases not requiring a firm power supply with as high a reliability factor as can be obtained with a double circuit line.

Controlled intermediate synchronous condenser compensation should be studied also along with the use of series capacitor compensation and it is hoped that Hoard will extend his studies to include this type of compensation too.

Sherwin H. Wright (Ebasco Services, Incorporated, New York, N. Y.): This paper on a series-capacitor-compensated transmission line is interesting and suggests some questions, the answers to which, I believe, will add both interest and value.

This paper gives operating results to be expected on a particular system, providing certain assumed conditions can be satisfied by development of special apparatus. If apparatus development is successful to the extent that the new problems encountered in series-capacitor compensation cannot merely be solved but can be solved economically, then the data in the paper become pertinent from a practical application viewpoint.

There are thus two aspects to the solution of these new problems:

(1) Can they be solved?
(2) Can they be solved economically?

This leads to the question, "Can Mr. Hoard describe the costs or relative costs on which some of his conclusions are based?" For example, in the conclusions, it is stated that "building the capacitors to withstand the high voltages during short circuits in most cases cannot be justified," but that in some systems "even though it may be a more expensive installation, may be economically feasible where firm power is a consideration."

B. V. Hoard: Evans' discussion supplies some of the important history and background regarding the development of series-capacitor compensation for long a-c lines. He states that it may be feasible to use other relaying schemes besides the current-balance scheme assumed in the paper. I of course would like to see a relaying scheme developed for series-compensated systems. Such a scheme should produce high-speed simultaneous tripping of a faulted line under the severe operating conditions where maximum load current is often greater than minimum 3-phase fault current. We also believe that an improved protective scheme can be developed and incorporated into the series-capacitor equipment which will reinsert them into the circuit more rapidly than 3.5 cycles after the fault is cleared, as was assumed in the paper. One possibility is an adaptation of the "Marx' High-Pressure Arc Valve," which originally was developed in Europe for high-voltage, high-power rectification and inversion uses.

A part of Wright's discussion leads me to believe that I did not explain sufficiently some of the assumptions used and conclusions reached in the paper. The system selected in the study, including a fault clearing time of 3.5 cycles at the near end and 7.0 cycles at the far end of the faulted line, with an additional time of 3.5 cycles for reinserting the capacitors after the fault is cleared, was selected as obtainable from a practical standpoint with equipment now available. However, I endeavored to point out that if relaying equipment and capacitor control equipment could be developed which would produce high-speed simultaneous clearance of the fault followed by immediate reinsertion of the capacitors, then a considerably lower value of compensation would be needed for the same transmission capability. The objective of development of these new equipments would be to obtain a system with less compensation required and lower cost for the same transmission capability.

With reference to a system design which does not remove capacitors during faults, in the body of the paper I state that if manufacturers allow twice rated voltage on the capacitors during the period of the fault, then a system having maximum fault current through the capacitors of 2.2 to 2.4 times maximum load current may use this design economically in comparison with the design in which the capacitors are short-circuited during the fault. In general, this condition can be approached if the per cent compensation is low and if the terminal equipment impedance is not too low. A somewhat higher voltage than 2.0 times normal can be built economically into the non-short-circuiting capacitors for the fault condition, because the system which short-circuits the capacitors during the fault when the voltage is greater than two times normal requires a higher per cent compensation, that is, a greater capacity impedance to obtain identical transmission capability.

I feel that the question of relative economics is pertinent but was considered outside the scope of my paper and would be too long and involved to be discussed here. This question is of major importance, and it is hoped that in the near future a member of the Bonneville Power Administration staff will be in a position to present a paper on this involved subject which will make economic and cost comparisons of compensated and uncompensated systems.

I agree with Crary in his discussion that somewhat different conclusions as to the importance of the various factors would be obtained if the studies had been based on specified system requirements of simultaneous fault clearance and upon immediate reinsertion of the series capacitors. In fact, I feel that these requirements should be perhaps the most important objectives of a series-compensated system design, and that equipment should be developed to meet these requirements. However, in making the studies, since we are engineers of an operating agency, we considered it necessary to assume a practical system which could be built and operated using the best existing equipment available. When the manufacturers can assure us that equipment has been developed which will produce simultaneous clearing of a faulted line and then practically immediate reinsertion of the series capacitors, we shall be glad to incorporate this equipment into the design because of the improved transmission capability resulting for the same percentage of compensation.

Three major objectives were in mind when this paper was written

1. To present the new operating problems produced by series compensation.

2. To show that these problems could not be solved fully by existing equipment available to the extent that something less than maximum transmission capability would have to be used until new equipment could be developed by the manufacturers.

3. To present sufficient data from studies to show that there is a real advantage resulting if this new equipment is developed and to present data regarding a compensated system's behavior during faults, including fault currents, voltages, and phase angles, so that the manufacturer's engineers could attack this problem without delay.

With respect to Crary's reference to additional improvement in capability through use of lower reactance and higher inertia generators, I agree in principle with the statement. It will be observed that a value of 33-per-cent transient reactance for generators was used in the studies, which is somewhat lower than for standard generators. However, depending on the specific problem, there may be some occasions where equivalent transmission capability can be obtained with lower cost by using a higher percentage of series compensation with standard generators as compared to the cost involved if lower reactance generators and lower percentage compensation are utilized. With the higher generator reactance of a standard generator, somewhat lower circuit-breaker duty would be obtained for faults at most locations on the system.

Sensitive Ground Relay Protection for Complex Distribution Circuits

Discussion and author's closure of paper 46-200 by Lloyd F. Hunt, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 765–8.

D. I. Cone (The Pacific Telephone and Telegraph Company, San Francisco, Calif.): The persistent and successful labors of Lloyd F. Hunt described in this paper and in its predecessor[1] constitute an inspiring example of search for the best engineering answer to a problem where many variant interests are involved. It is a specific example of that professional breadth of view which engineers should seek to exemplify.

REFERENCE

1. SENSITIVE GROUND PROTECTION FOR RADIAL DISTRIBUTION FEEDERS, Lloyd F. Hunt, J. H. Vivian. AIEE TRANSACTIONS, volume 59, 1940, February section, pages 84–7.

M. A. Sawyer (Southern California Telephone Company, Los Angeles, Calif.): As noted in the paper, the ground protection system provides not only a means of promptly de-energizing a feeder in case of wire failure involving ground contact even where the contact resistance is relatively high, but it also affords an effective means of minimizing the hazard involved in the joint use of poles by higher voltage power circuits and communication facilities. Experience has shown that the amount of damage to communication plant and the hazard involved in cases of contacts by power wires are to a large extent a function of the duration of the contact and the amount of fault current. Where higher voltage joint-use is employed it is therefore important that these factors be made as small as practicable. The ability of the relay system, described by Lloyd C. Hunt, to limit the ground fault current to a low value and at the same time provide for the prompt tripping of the faulted circuits makes it particularly well-suited for such situations.

General joint use of telephone facilities and 7-kv distribution feeders equipped with sensitive ground protection was entered into in an industrial area near Los Angeles in 1937. Approximately 700 jointly used poles were involved initially, and the number has increased now to about 850. The power distribution system consists of radial feeders with no looping of feeders, and the telephone facilities are almost entirely in cable. In 1940, sensitive ground protection was installed in a semirural area in Orange County which facilitated rather general joint use of 11-kv distribution circuits and open wire as well as cable telephone facilities. Looped and radial feeders, as indicated by Figure 2 of the paper, as well as radial feeders from a source station are involved. There are at present about 1,600 jointly used poles in that area. A small amount of joint use with 11- and 16-kv feeders is employed also in two other areas. The present total number of jointly used poles with 7-, 11-, and 16-kv circuits equipped with sensitive ground protection is therefore approximately 2,500.

During the past eight years, twelve contacts between supply feeders equipped with sensitive ground protection and telephone plant occurred, the majority of which were involved at crossings rather than in joint use sections. In all cases the supply circuits were de-energized promptly and the telephone protective devices operated properly. In general, the amount of damage to the telephone plant and the resulting service interruptions were rather minor. Experience to date therefore indicates that the relay system provides a reliable means of limiting the adverse effects of accidental contacts.

Considerable effort is being directed toward the extension of both power and telephone services into the rural areas where all possible economy in line construction must be realized. Joint use of poles by power and telephone facilities is one means of reducing the cost to both parties of such construction provided the necessary safety can be obtained. Where power systems having multigrounded neutrals are involved, a satisfactory degree of protection co-ordination usually can be obtained with the protection normally employed in the power system by the application of auxiliary protectors on the telephone lines, using the power neutral as the grounding medium. Where a power circuit having only the usual relay and fuse protection does not have a multigrounded neutral, joint use is generally undesirable because low impedance grounds for the telephone protectors are not normally available in rural areas and protection co-ordination cannot be obtained. However, the sensitive ground protection system, described by Hunt, should provide a high degree of safety in such situations because it is not essential that the impedance to ground of the telephone protectors be low to insure tripping of the faulted circuit in case of contacts.

Also, with the high impedance in the small power system grounding banks, the impedance to ground in the telephone system can be relatively high without producing voltages on the telephone plant which will exceed the safe operating limits of the telephone protective devices.

Hunt is to be complimented on the development and application of a relay system which not only offers a practical solution to the problem of clearing ground faults on complex as well as simple distribution circuits, but also affords an effective means of providing a high degree of safety where joint use of poles by higher voltage distribution circuits and communication facilities may be desirable.

F. A. Cowan (American Telephone and Telegraph Company, New York, N. Y.): Lloyd F. Hunt's paper describes a power relaying system which is of considerable interest from the standpoint of limiting the adverse effects of a contact between power and telephone systems under certain conditions.

Experience over many years indicates that a factor of primary importance in promoting a high degree of safety in situations where power and telephone systems are in proximity is the speed and reliability with which the power current is interrupted if a contact occurs between an energized power wire and a telephone cable or open wire circuit. This is true whether the power and telephone circuits are on the same pole line or are involved in numerous crossings.

With multigrounded or common neutral power circuits, prompt and reliable interruption of power current in the event of a contact is usually relatively easy to secure. The neutral is an extensive well-grounded conductor which is readily available so that the telephone plant can, if desired, be connected to it at frequent intervals, directly in the case of telephone cable sheaths and through protector gaps in the case of telephone open wire. Under these conditions, the current available for operating power circuit fuses or relays in the case of a contact to telephone cable is not much less than that available in the case of phase-to-neutral short circuit, and de-energization is almost always prompt and reliable. Where the telephone circuits are open wire, similar approximate equivalence of contact current and phase-to-neutral short-circuit current can be obtained if the voltage to ground of the power circuit is above the operating voltage of the protector gaps which, at the present state of the art, is about 3 kv. Where joint pole lines are used, the connections to neutral usually can be made easily by connecting the cable sheaths or protector ground leads to vertical power ground wires; where separate lines are used the same method can be applied if joint crossing poles are used extensively.

However, where the power circuit does not have a multigrounded or common neutral, situations sometimes arise where the impedance presented by the telephone plant in the event of a contact is high enough to reduce significantly the fault current as compared to that in the event of a phase-to-neutral short circuit. For example, in some rural areas where well-grounded power secondary neutrals are not available, reliance must be placed on made grounds for grounding telephone cable sheaths and protectors and the resistance of these grounds under adverse conditions may be relatively high. In occasional cases of this sort, the current available in the event of a contact will be so limited as to cause relatively slow de-energization of power relaying and fusing systems of the conventional types.

As a result, any development such as that described by Hunt which tends to reduce the magnitude of ground fault current necessary for reliable and prompt de-energization of power circuits without multigrounded or common neutrals, tends also to facilitate the provision of a high degree of safety where joint use or numerous crossings with such power circuits exist. The telephone engineers will, therefore, I am sure, welcome the use of Hunt's scheme in those cases where its application is advantageous from the power standpoint.

Lloyd F. Hunt: The discussions presented by F. A. Cowan, M. A. Sawyer, and D. I. Cone are appreciated very much. They point out the value of mutual study on problems that are confronted by both types of utilities to render good service to the common users. It can be understood from their discussion that the small added cost for sensitive ground protection on high voltage joint use of pole lines is more than justified and is very much to the advantage of the power company if required to take alternate methods of serving, such as underground or entirely new right-of-ways.

An Electronic Frequency Meter and Speed Regulator

Discussion and author's closure of paper 46-188 by Ellis Levin, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, December section, pages 779–86.

Godfrey T. Coate (nonmember; Massachusetts Institute of Technology, Cambridge, Mass.): Ellis Levin's paper describes an ingenious way of employing electronic circuits to make an unusual speed-measuring device. Without the aid of phototubes and amplifier tubes it certainly would be difficult, if not entirely impossible, to measure a high speed of rotation with such convenience and accuracy and without placing a mechanical load on the rotating shaft. I have read the paper with interest and should like to comment on two details of the circuits described.

In the derivation of equation 5 of the paper (the relation between frequency and the resistance R_K), reference is made to the circuit of Figure 3, and an approximation is required in obtaining the final equation. This approximation may be avoided if the derivation is made for the circuit used in the actual instrument. For this circuit, shown in Figure 4, the voltage across R_B is $2fCIRR_B$ (see equation 1) exactly as in the circuit of Figure 3, but the voltage across R_K is IR_K instead of the expression given in equation 2. Adjustment of R_K so that these two voltages have equal magnitudes then yields $2fCIRR_B = IR_K$, and from this expression equation 5 may be obtained.

The feed-back circuit used in the 6L6 amplifier of Figure 8 is interesting in that it seems to be very different from ordinary negative feedback. The 6L6 tubes operate from a 60-cycle-per-second plate voltage supply and receive a 60-cycle-per-second signal. The output of the stage is the difference of the average components of the two tube currents, and this difference is roughly proportional to the amplitude of the input signal. The feed-back voltage is not porportional to the output, but rather to the sum of the two tube currents, and its wave form is probably a series of half-sine-wave positive pulses. I never have seen an analysis of this type of feedback and should like to know more about its effect upon the amplification and stability of the circuit.

Ellis Levin: Unfortunately, the circuit of Figure 4 of the paper does not yield the exact linear equation of balance that G. T. Coate suggests, although it may appear to do so at first glance. Perhaps Coate reasoned as follows: Consider Figure 4 of the paper, but with half of the switching circuit removed; that is, eliminate for the time being V_3, R_{L3}, C_3, V_6, and V_7, as shown in Figure 1 of this closure. Then in this simplified circuit the resistor R_B conducts only capacitor discharging pulses, while the current in R_K is composed of capacitor charging pulses plus capacitor discharging pulses plus a steady component. In R_K the charging and discharging pulses flow in opposite directions, and therefore it might appear that these two transients cancel one another, leaving only the steady component and making the voltage across R_K independent of frequency. Also it might appear at first glance that the current in R_B is exactly proportional to frequency, so that the setting of R_K required to balance the output voltage to zero would be exactly proportional to frequency. Actually this is not strictly correct, for two reasons. First of all, the two transients in R_K have unequal magnitudes and therefore do not exactly cancel one another. Secondly, as a result of the location of R_K, the voltage across R_B is not strictly independent of the setting of R_K.

A closer inspection of the circuit will show that the voltage across R_B (in Figure 1 of this closure) is not $2fCIRR_B$ but rather

$$V_{RB} = fCI(R_L + R_K)R_B \qquad (1)$$

and the voltage across R_K is not IR_K but rather

$$V_{RK} = \frac{IR_K}{2} + fCI(R_L + R_K)R_K - fCI \times (R_L + R_K)\frac{R_P R_K}{R_P + R_L + R_K} \qquad (2)$$

Equating these two voltages of equations 1 and 2 of this closure and solving for f yields

$$f = \frac{R_K}{2C(R_K + R_L)} \times \frac{1}{\dfrac{R_P R_K}{R_L + R_P + R_K} + (R_B - R_K)} \qquad (3)$$

From equation 3, f is not exactly linear with R_K. If we plot f versus R_K for frequencies from zero to 833 cycles per second (which corresponds to a turbine speed of 100,000

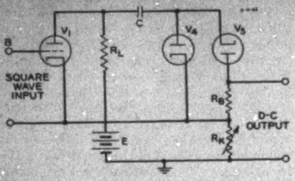

Figure 1. Circuit of Figure 4 of the paper with half of the switching circuit removed

rpm, as the input trigger amplifier halves the turbine frequency) and draw a straight line between the two end points, then the maximum deviation of the curve from the straight line will occur at approximately the mid-frequency, and will be equal to about 5 per cent of this frequency or 2.5 per cent of full scale.

Equation 3 of this closure was derived for the simplified circuit of Figure 1 of this closure. If we add the other half of the switching circuit to obtain Figure 4 of the paper we cannot say that the equation of balance for Figure 4 is exactly the same as equation 3 of the closure because the configuration of Figure 4 provides a larger number of paths for the capacitor charging and discharging currents. An exact solution of Figure 4 involves at least six different branch currents at each switching instant, and becomes rather long and cumbersome. The actual calibration curve of Figure 4, determined experimentally, is very similar to that given by equation 3 of the closure. Its maximum departure from linearity is about five per cent of full scale. However it is *accurate* to better than one-half per cent at any frequency from 83.3 to 833 cycles per second corresponding to turbine speeds of 10,000 to 100,000 rpm. By altering the circuit constants, the same accuracy can be achieved for other frequency ranges.

Specific Engineering Problems in Rural Electrification and Electroagriculture

Discussion and author's closure of paper 46-184 by M. M. Samuels, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1065–73.

Hamilton Treadway (Rural Electrification Administration, Ashton, Md.): The lack of poles is among the problems pressing the industry at this time. The development of new methods of treatment is desirable. Effective treatments which can be applied quickly are needed to speed the delivery of poles. For those new treatments which have been introduced survivor life data are needed. This is a field in which practical methods of predetermining the average life are almost wholly lacking and performance characteristics of new treatments must be secured principally through field trials. The development of some practical testing procedures which will permit a reasonably accurate prediction of pole life will stimulate the development of new treatments and techniques.

Has there been any experimental work on the use of silicone resins in the treatment of poles and crossarms? If effective as a pole treatment, how do costs compare with existing methods? Would the use of silicones improve the insulation characteristics?

Another approach to the pole problem lies in a study of the average life of treated and untreated poles, comparing them on a basis of pole strength remaining at the end of 10, 15, and 20 years. A study of this nature will show whether it is more economical to invest in wood or in treatment. Are any data available on this subject?

The author suggests an interesting problem in proposing a statistical study of load diversity. Hamilton[1,2] and others[3,4] have shown that the load duration curve can be represented mathematically by a probability curve. In general, the characteristics of these curves are determined by the load factor and the ratio of the maximum demand to the minimum demand. The evaluation of these factors for the various types of rural loads will make possible an analysis of the economics of many factors affecting the design of lines and equipment.

The development of an inexpensive load factor meter with an attachment to record the minimum demand is a possibility. The maximum demand and the diversity factor can be determined if the load factor and kilowatt-hours are known.

The economics of voltage regulation is given also to statistical study. If the ordinates of a voltmeter chart are rearranged in descending order, they also approximate a probability curve. Here again the average voltage and the maximum and minimum values determine the characteristics of the curve. Could a simple instrument be developed to record these values? This is a different approach to the problem of voltage regulation and economics and would entail new methods of analysis. The effect of voltage regulation is an important factor in the economical operation of equipment, such as transformers, and the approach suggested here may simplify the evaluation of the economics of voltage regulation.

New industries have developed in the rural areas as power facilities have been extended. Much of the development of rural lines has been conservative, and capacity is usually lacking to serve these industries. Here again enough is not known about load characteristics. Extensive field tests are necessary to determine such factors as duty cycle, voltage and load swings, light flicker, as well as load factor, diversity factor, power factor, and minimum and maximum demands. Once assembled, these data will be useful in the design and operation of the power facilities and in improving the operating characteristics of the consumer's equipment and processing methods.

The improvement of the transformer is not static. In addition to reducing losses and weight an increase in the insulation level is desirable. The use of class C insulation with the new silicone binders is suggested. These insulations offer the possibility of higher operating temperatures. This may be the key to the problem of reducing core losses. By operating the core at a higher temperature it is possible to take advantage of MacLaren's effect.[5] This can be accomplished by the insertion of heat barriers. At present, further increase in operating temperatures may be limited by sludging of the oil. The development of a new cooling medium is necessary. The possibility of high pressure gas is suggested. The difficulties encountered in its application while still unsolved are not insurmountable. High pressure gas cooling also holds out the possibility of some further reduction in weight.

What is the possibility of reducing losses by using powdered core materials? Can the permeability of such materials be increased to a point where they will compare favorably with other core materials for power transformers?

The progress of our civilization has been made possible by the free exchange of technical information and ideas. The author has suggested a multitude of engineering problems which requires the application of our best efforts. It is to be hoped that free discussion of these problems will bring out many new ideas to improve both rural electrification and electroagriculture.

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3. Evaluating Diversity Factor, A Statistical Method, B. D. Price. *Electrical Review*, London, England, September 25, 1936, page 400.

4. Diversity and Load Factor, D. J. Bolton. *Electrical Review*, London, England, April 17, 1936, page 563.

5. The Effect of Temperature Upon the Hysteresis Loss in Sheet Steel, Malcolm MacLaren. AIEE Transactions, volume 30, pages 761–75; volume 31, pages 2025–35.

J. R. Werth (Department of Agriculture, Washington, D. C.): The author stresses the importance of continuity of service to the rural user. Demand charges promote a single source of wholesale supply and hence are detrimental to continuity. A plea is made for studies tending to eliminate demand charges.

This elimination should not be difficult to justify. Frequently the farmer co-operatives furnish the substations. Two or more sources of supply inevitably will reduce interruptions and will promote continuity. Regardless of its prior justification, today a demand charge is morally equivalent to a cover charge in a night club. You pay something for nothing. Would not any self-respecting game hen of American or of Indian descent be outraged if the egg buyer paid the farmer's wife a demand charge for readiness to lay and also an energy charge for eggs? Admit that in the remote past eggs and kilowatt-hours could not be stored readily. The modern locker plant and pondage back of the hydro plant tied into an electric grid now do store eggs and kilowatt-hours. Therefore, I present this plea that in the interest of continuity demand charges be declared out.

Why not (as John Carmody said when cyclometer dials were suggested for rural kilowatt-hour meters in 1937)?

Where? First, in most advantageous locations (as inventer, W. L. R. Emmet, invariably recommended for innovations).

When? Now, in 1947!

O. W. Gatchell (Frank Horton and Company, Lamar, Mo.): This discussion of Samuel's long-awaited paper is directed towards a simplification of the engineering design problem in the rural electrification field.

1. Use right-of-way easements that provide for possible future underground transmission or buried neutral construction, in acquiring present distribution right-of-way. This will simplify the future designer's problem.

2. A "superman" rural breaker, even with a high electrical intelligence quotient of 140 plus, is no substitute for an axe. A definite tree-trimming maintenance program simplifies the breaker's job and means that sectionalizing equipment design can be brought down to earth.

3. Samuels mentions 7,200-volt circuits. Why not standardize now on 7,620/13,150 volt rating? That voltage is as high as that which can be used below the code 15 kv ceiling, where ground clearances change. Voltage of 7,620 instead of 7,200 means 12 per cent more line capacity for the same drop; distribution transformers are identical in cost. Where else can the designer give the farmer 12 per cent more capacity for negligible outlay? All that is needed is a program to educate the power transformer manufacturer.

4. An inviting panorama of greatly increased rural use of electricity is unrolled in the author's paper. However, nothing is said about locating the transformer at the center of all of this load, between kitchen and barn. Farmstead location of the transformer will simplify present overhead and future underground secondary installation; it even could help simplify application of primary underground lead-in from the overhead line along the road, if such a development should prove to be the first practical use of single phase rural primary underground.

Max W. Rothpletz (Rural Electrification Administration, Washington, D. C.): In Table I of this discussion are some figures which portray the conditions existing in two rural distribution systems picked at random and analyzed on the basis of known data and conservative estimates.

These figures further emphasize the need for lower core losses in rural distribution transformers.

The most economical transformer design is one which provides the required quality of service for the lowest combined fixed and variable charges. Quality of service includes consideration of such items as regulation, telephone influence factor of the exciting current, and exciting current itself. On rural circuits the first two items should be maintained as low as possible, but the exciting current may be increased somewhat if necessary to gain some other desirable feature. In accordance with the definition in the Federal Power Commission glossary, fixed charges include interest, depreciation, property taxes, and property insurance. All other charges are variable charges.

Because only minor increases in transformer efficiencies can be anticipated, transformer losses must remain at about the present values, but the ratio of core to copper losses can be selected so that the total energy losses over a period of time are lower.

Table I

Type of Loss	System A	System B
Distribution transformers		
Core losses, per cent	64.3	60.3
Copper losses, per cent	9.2	9.2
Subtotal, per cent	73.5	69.5
Substation transformers		
Core losses, per cent	6.4	7.0
Copper losses, per cent	6.1	5.2
Subtotal, per cent	12.5	12.2
Meter iron losses, per cent	4.2	3.3
Line and miscellaneous losses, per cent	9.8	15.0
Total, per cent	100.0	100.0
Total losses in per cent of purchased power	24.8	21.3
Miles of line	525	1,025
Number of consumers	1,300	2,400

However, it may not be practical to design for the lowest energy losses because the cost of equipment may be increased to the extent that fixed charges are increased more than the variable charges are reduced. A formula has been derived to show the relationship between core and copper losses and the additional money which may be spent on equipment to obtain the desired characteristics.

Since copper losses vary as the square of the transformer load, the total energy losses depend on the load curve. If instantaneous values of copper losses are calculated for a particular load curve, a loss curve may be drawn from which a loss factor showing the ratio of average losses to peak losses may be derived. Then if

L is the loss factor in per cent
R is the energy cost in cents per kilowatt-hour
F is the fixed charges in per cent
C is the decrease in core losses in watts
W is the number of watts of increased copper losses at peak load

the permissible additional investment in equipment, P, is given by

$$P = \left(C - \frac{LW}{100}\right) 8.76 \left(\frac{R}{F}\right) \text{ dollars}$$

For demonstration purposes an assumed transformer load curve was analyzed and found to have a loss factor, L, of 10.9 per cent. Assuming, for simplicity purposes, that $R = \$0.009$ and $F =$ ten per cent, an additional transformer investment of \$0.70 would be permissible if the core losses could be decreased one watt with an increase of one watt of copper losses at peak load. If the design required that the copper losses be increased two watts at peak load the permissible additional investment would be \$0.61.

When a transformer load curve is not available for determining the loss factor, the minimum permissible additional investment can be calculated by the same formula using the load factor. The load factor of the transformer used in the aforementioned example was found to be 23.4 per cent. Using the same assumptions as mentioned previously, a minimum additional investment of \$0.60 would be permissible if the core losses were decreased by one watt with an increase of one watt of copper losses at peak load. If copper losses were increased two watts for one watt decrease in core losses the minimum permissible additional investment would be \$0.42.

In selecting a transformer size for an individual load, it is important to make sure that it will not be too large for a long time to come, not only because of the first cost but also because of the core losses.

For example, a one kva transformer is estimated to cost \$34.00. A $1\frac{1}{2}$ kva conventional transformer costs approximately \$39.00. If the load is not expected to increase over and above one kva, the $1\frac{1}{2}$ kva size would represent an unnecessary investment of \$5.00. Furthermore, the losses of a one kva transformer are about 10 watts core and about 50 watts copper; at 1 kva load the losses of a $1\frac{1}{2}$ kva unit are about 21 watts core and 20.5 watts copper. Based on the same assumption as in the previous example, using the one kva transformer the saving in losses would be equivalent to saving a further additional investment of \$6.13. If the savings in losses had been calculated on the basis of the load factor of 23.4 per cent a minimum savings in investment of \$3.24 would have been indicated.

M. M. Samuels: The discussions were somewhat disappointing because of the lack of more constructive ideas. Rothpletz's elaboration on the core loss problem is very much to the point. A suggestion may be added for the rate man. At present the core losses are billed wholesale under the most expensive demand block and the most expensive energy block. Should not the core losses be deducted from the total demand for billing purposes and should not the energy be billed at the lowest rate per kilowatt-hour?

Treadway raises the question whether or not anyone has experimented with the use of silicone resins for the treatment of poles and crossarms. An inquiry among specialists in this field did not reveal any such experiments so far. Those who had any experience with attempts to make demand diversity studies will all agree that Treadway's proposed method will not solve the problem. Such a study, to be of any value at all, would call for an investment of great sums in recording meters. It would have to be carried out over long periods and diversified in difficult climates. After spending all that money on survey and analysis, the conclusions, if any, would be very likely obsolete by the time the work would be finished. Treadway's other suggestions are recommended for consideration by manufacturers. No doubt manufacturers already have studied these features.

Gatchell suggests that rural circuits be standardized at 7,620/13,150 volts. When big time construction of rural circuits began, it was found that the price of equipment, particularly transformers, jumped considerably at the 12,500/7,200-volt level. At that time no one predicted that the load would grow as it actually has grown. This voltage was then the most economical. In fact, many engineers were of the opinion that it was too high a voltage and suggested continuing the use of 2,300 volts which was then the most common voltage of rural circuits. Now, it would be difficult to change the voltage of hundreds of thousands of miles of line. The problem of improving the rural circuit is being studied everywhere now, and the studies will result undoubtedly in great improvement everywhere. Consideration is being given to additional sources of power at various places on a rural system, backbone lines of higher voltages, series capacitors for voltage regulation, and other methods.

As the cost of overhead construction increases, the difference in cost between overhead and underground may become less pronounced. Cable manufacturers are engaged now in elaborate studies of possible underground primaries for rural circuits. Gatchell's other suggestions deserve serious consideration.

Silicone-Resin-Treated Magnet Coils

Discussion and authors' closure of paper 46-104 by Graham Lee Moses and Julius J. Torok, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 412–16.

B. W. Erikson (General Electric Company, Schenectady, N. Y.): The authors have introduced what they call the *K* factor. We agree that this is a very useful method of evaluating coil deterioration under certain conditions. However, we would like to point out that a consideration of this *K* factor may be misleading when considering designs involving different types of insulation. For instance, we have found that on conventional class *A* insulated coil structures at normal operating levels there is often no appreciable change in the *K* factor even when the coil has fallen apart completely as a result of prolonged operation at high temperature.

However, and this is significant, all good insulation systems insulated with silicone, or possibly with high temperature synthetics, show a marked change in the *K* factor with temperature-time aging. This may be taken to indicate that where the coil impregnant contributes materially to the original excellent heat transfer rate of the system an ultimate large change in the *K* factor should be expected. Much work remains to be done in exploring and evaluating this relationship.

With reference to the section entitled "Effect of Operating Temperature on Design," the authors refer to Figure 8 and indicate that this shows ampere turns at various levels of coil temperature rise. It seems to me that this is too general and perhaps not directly applicable to a large variety of d-c contactor coil designs similar in nature to those coils referred to in this paper.

To illustrate, I made a comparison some years ago between the ampere turn gain or loss occasioned by changing a design from either enamel wire insulation or enamel-cotton insulation to either glass or asbestos insulation. It became evident from this survey that there is no gain whatever in the finer wire size. Comparing asbestos and double glass with enamel there was found to be no ampere turn gain on designs using wires smaller than approximately 0.075-inch diameter.

In making this same comparison with enamel-cotton insulated wires there was found to be no appreciable gain below about 0.040-inch diameter. Included in this comparison was the fact that the class *B* designs were operated at a temperature 20 degrees higher than the class *A*. This confirms the authors' statement that there is little if any gain in the use of conventional class *B* designs as compared with class *A* unless the temperature level can be increased materially above the 145-degree-centigrade total of class *B*.

Where these facts are applied to conventional d-c contactor and relay designs, it is difficult to see where any gain in ampere turns or size of coil is to be made. Common 125- and 250-volt d-c coils such as those discussed in this paper, use wire sizes ranging from approximately 0.007 up to and including 0.014-inch in diameter. Comparing the space requirements of these small diameter wires between those insulated with conventional enamel and those insulated with either single or double glass, we find that 0.010-inch-diameter wire if insulated with single glass requires 168 per cent of the space taken by enamel, and 0.014-inch-diameter wire if insulated with single glass requires 145 per cent of the enamel wire space. On this basis the silicone coils would

1. Produce for the same ampere turns approximately 50 per cent more watts.

2. Operate at a temperature approximately 50 per cent higher merely to break even.

If I interpret this correctly, it would mean that silicone designs would have to operate in the neighborhood of a 125-degrees-centigrade rise (this would be the class-*A* 85-degree rise plus 50 per cent equals 125 degrees centigrade) so as to compare in performance with conventional class *A*.

Any higher operating limit would produce a gain in ampere turns. I believe that this particular point, even though contained between the lines in the paper, needs some additional emphasis and further consideration by design engineers.

Carl J. Fechheimer (consulting engineer, Milwaukee, Wis.): There are a few points to which I would call attention.

In regard to the dissipation of heat, it is of interest what part of the temperature drop is inside the coil and what part is at the surface. Also, what are the maximum internal temperatures?

It generally is recognized that there is a limit of stability when coils are operated at high temperatures when the current is kept constant. For as the temperature rises, the electrical resistance increases, and therefore the I^2R loss increases. This in turn raises the temperature further, and at some critical value the temperature will rise without limit. As the heat is dissipated from the surface by radiation and natural convection, both of which increase more rapidly than the temperature, these effects offset somewhat the effect caused by increase in resistance and raise the stability point. As 2-dimensional heat flow and heat storage must be considered, the problem becomes very complex. Have the authors investigated the thermal stability of the coils, either experimentally or theoretically?

The mounting resistance with temperature also means that before the stability point is reached, the most economical point of operation is passed from the standpoint of the materials and from that of the power consumed. Have the authors made an investigation of the most economical point?

If the coils are operated in shunt with the supply line, the ampere turns decrease as the temperature increases. Then the I^2R decreases with mounting temperature, and the question of stability does not enter. Or do the authors propose to increase the applied voltage so as to maintain constant ampere turns? If not, if the pull of the magnet is proportional to the square of the current, the pull of a shunt magnet coil is approximately only 36 per cent at 200 degrees centigrade of that which it is at 25 degrees centigrade. Is that not a serious decrease?

L. A. Doggett (Pennsylvania State College, State College, Pa.): Improvement of magnet coil design is important because of the millions of such coils in industrial use. To meet specifications with an ever decreasing amount of active material has been the long standing objective of the design engineer. The new silicone insulations provide a long step in this direction.

The reduction in copper weight as the temperature rise is allowed to increase is shown in Figures 9 and 10. Not all of this gain, however, may be chargeable directly to the real advantages in the use of silicone insulations. If it is assumed that temperature rises of 160 degrees centigrade may be used safely with silicone insulation, then to provide the ampere turns specified in these coils the length of the coil may be reduced about 19 per cent keeping the outside diameter at three inches. Similarly if the coil length be kept at $3^1/_4$ inches, the coil diameter may be reduced about 15 per cent. The weights in these two cases are reduced correspondingly some 19 and 38 per cent. Coincidental with these size reductions is an increase in the ratio of total coil surface to coil volume of 9 and 23 per cent. For many years transformer designers have been familiar with the general fact that the smaller the transformer volume the greater is the ratio of surface to volume. For example, a cube with a 2-foot side has a surface-to-volume ratio of three to one, whereas a cube with a 1-foot side has a surface-to-volume ratio of six to one.

Summarizing, it may be said that by using silicone insulations, coils may be made smaller, and because they are smaller they can be worked harder.

J. M. Fluke (Anderson Fluke Engineering Company, Springdale, Conn.): The authors are to be complimented for this paper and the extensive amount of test work which has gone into the data presented. Without question the silicone-glass insulating materials represent a very marked improvement over materials previously available. The United States Navy Bureau of Ships has made an effort to exploit the use of these new materials as fast as they have become available and to carry their application to the maximum extent possible for the immediate purpose of learning more about their properties and for the ultimate purpose of increased reliability and weight and space saving. With the co-operation of industry there are on order equipments and development programs covering nearly every type of wound electrical apparatus ranging from small motors through generators as large as 1,250 kva at 450 volts alternating current, transformers, and welding sets. In addition, the use of silicone-glass materials in producing highly arc resistant and stable sheet and molded insulating material is receiving concentrated attention.

Present tentative proposals provide for operating temperatures of silicone-glass insulated electrical machinery as high as 200 degrees centigrade total temperature. Weight savings on transformers run as high as 30 per cent with savings approaching this value on rotating apparatus.

It is apparent then that this new insulation offers appreciable advantage to the electrical plant in naval vessels in view of the ever increasing necessity of weight reduction. In addition, as these materials improve in quality as a result of the continued effort toward improvement of their characteristics and methods of application, materially increased reliability of the apparatus in which they are used will be realized. In fact it is safe to say that the improvement in this respect is sufficient to warrant intensive interest if no weight or space saving were involved at all.

In specific comment on this paper, several points are worthy of additional mention. Under the tentative temperature limits now being used, a 40-degree-centigrade allowance is made for hot spot temperature when measurements of temperature of a silicone-glass insulated coil are made by thermometer, and a 20-degree-centigrade differential is made when resistance determination of temperature is made. It is felt that these figures require additional investigation. It is quite probable that for many applications they are too large. In particular it is suspected that this is true for magnet coils of normal size. This statement is based on measurements made some time ago with thermocouples placed in coils operated at class *A* insulation temperature where the maximum observed temperature differential through the coil did not exceed several degrees centigrade. A much larger allowance previously made for this condition could be reduced through this knowledge which in turn effected some saving of weight and material so necessary during wartime construction. It is recommended therefore that future life tests of silicone-glass insulated devices incorporate suitably placed thermocouples for obtaining more data in this respect. This information will be particularly pertinent to the "Conclusions and Recommendations" given by the authors in their paper.

It is noted, but not conclusively, that a suitable means of preventing the oxidizing of copper may be a necessity in the ultimate satisfactory application of the silicone materials. Several premature failures of equipment have been traced to the possibility of failure because of the reduction of copper cross section from oxidation. Additional test work is necessary to establish this statement definitely. However, the continued development of silicones may allow even higher operating temperatures. To meet this possibility there is need for further investigation of this condition and the development of suitable means for its prevention by plating, preferably, as the authors suggest.

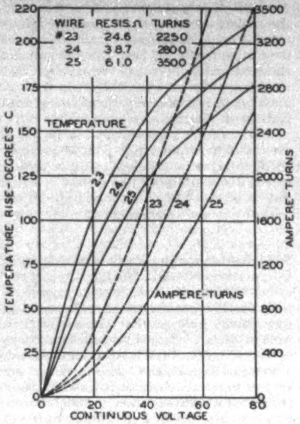

Figure 1

In conclusion, it is considered that the availability of silicone-glass insulation is one of the most significant insulation developments since the art began. Its widespread use in the not too distant future is a certainty. The versatile silicones are only in their comparative infancy. So far they have been employed successfully as varnishes, fluids, and resins for producing all types of electrical insulation. Combined with glass cloth or fibers, they produce highly stable molded and sheet insulation. The silicones also have performed excellently as lubricants at elevated temperatures. Their full application possibilities still are to be realized.

Graham Lee Moses and Julius J. Torok: The coils reported were shunt windings. Therefore no checks were made to determine "run-away" temperature conditions such as Carl J. Fechheimer suggests. It would be desirable to make such tests on series coils with silicone insulation.

No determination was made of temperature distribution on the coils reported. However, on similar coils it has been found that the average surface temperature rise is approximately 60 per cent of the internal rise by resistance and that the hot spot is within ten per cent of the average rise by resistance.

The change in the *K* factor is a useful means for observing aging where it occurs. It is recognized that this change in *K* does not occur necessarily on all types of coils, particularly on coils which have little or no varnish fill. As indicated by B. W. Erikson, much work remains to be done in exploring and evaluating this relationship. It is hoped that other investigators will study this problem.

The curve in Figure 8 shows the effect of operating a specific design of coil (the one used for these tests) on various voltages which produce different temperature rises. It shows the resultant hot ampere turns after temperature stability is reached. To supplement this a curve has been prepared showing actual values of temperature rise and ampere turns as a function of operating voltage for three different designs for the same spool (differing only in wire size). This curve shows that a coil wound with number 25 wire rates 52.5 volts at 85 degrees centigrade rise (by resistance) and produces 2,270 ampere turns (hot). The corresponding design of coil with number 24 wire would have a temperature rise by resistance of 121 degrees centigrade when operated at the same voltage and would produce 2,580 ampere turns (hot). Thus the increase in net ampere turns (hot) is 13.8 per cent. The increase in cold ampere turns is 26 per cent before the coils heat up to operating temperature. A design of coil with number 23 wire would operate at 170 degrees centigrade rise (by resistance) at the same voltage (52.5) and would produce 2,910 ampere turns (hot). This is an increase of 28 per cent over the coil with number 25 wire.

The lack of a silicone wire enamel is a distinct handicap in the development and application of silicone insulated magnet coils especially for finer wire sizes. It is recognized that glass wire covering has a poor space factor. The paper states that on finer wire sizes no gain may be possible. However, generalizations should be avoided and specific designs should be considered before deciding whether or not high temperature insulation offers advantages. It has been found possible to obtain an appreciable increase in ampere-turn rating with class *H* insulation on coils which use wire sizes of number 30 (0.0100 inch diameter) or even finer.

As stated by L. A. Doggett, the increase in ratio of surface to volume is a distinct advantage in heat dissipation. This was not included in the calculations made for this paper as it was assumed that all the heat was dissipated from the barrel surface. Therefore, in the size and weight advantage to be derived from high temperature, silicone insulation may be even greater than indicated by the curves shown in Figures 9 and 10.

Operating Experience With Distance Ground Relays

Discussion and author's closure of paper 46-83 by W. A. Wolfe, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, July section, pages 458–62. This paper also was presented at the South West District meeting, San Antonio, Tex., April 16–18, 1946.

Eric T. B. Gross (Illinois Institute of Technology, Chicago, Ill.): Though both reactance type relays and their proper connections for ground fault distance protection have been available for about 20 years, distance relaying has been applied in the past nearly exclusively for between phase faults, involving more than one conductor of the 3-phase system. W. A. Wolfe therefore is covering new ground, and he deserves commendation for his courage to try a new thing and congratulations on his success. It will be to the advantage of the art if engineers learn from this paper how easily satisfactory results can be achieved; after all, with the majority of faults being single line-to-ground faults (84 per cent in this case), it is evident that proper protection will provide great improvements.

Wolfe's analysis of the reasons for use of the distance ground relays are truly sound, and it is gratifying to see that the application of resonant grounding through ground fault neutralizers duly was considered. A voltage of 60 kv is definitely one at which ground fault neturalizers, preferably designed for continuous operation, could be applied very successfully.

The installation described in the paper, designed more than seven years ago, consists of three distance phase and three distance ground relays. It is possible to reduce the number of distance elements by proper control of the current and voltage connections of the relay scheme as described in references 3 and 4 of the paper. With the knowledge of our present day developments in the field of relaying, it should be easy to design a distance relay scheme, including ground fault relaying, which does the job with less than six distance elements and without increase in tripping time.

R. E. Cordray (General Electric Company, Philadelphia, Pa.): We believe that W. A. Wolfe has done the industry a great service in presenting his operating experience with ground distance relays, and we are particularly pleased to learn that he has ten terminals of *GCX* relays. I was pleased to be a member of the relay subcommittee which recommended that this paper be presented at the national meeting after it had been presented at the South West District meeting because I think that the subject matter warrants it.

We do not understand why ground distance relays have not been more popular. They have been in the catalogues of manufacturers for almost a decade, yet their use has been limited. The scheme as it stands now is as simple as phase relaying with the exception of the blocking units for preventing false operation on 2-phase-to-ground and phase-to-phase faults. Another exception of still less importance is providing for mutual compensation when it is required by parallel lines. The scheme should be kept simple, at least until ground distance relays become the accepted practice. This may be the answer to Doctor Gross's question about using one relay and switching it. We realize some possible excuse for expecting ground relays to be relatively simple and inexpensive as a result of the common use of a single overcurrent relay for ground protection. This has led relay engineers to believe that one relay should take care of all ground faults. When a directional ground relay was required, however, three potential transformers generally were required and the cost of the third potential transformer makes it about equal to the phase relaying cost.

Another possible influence in the use of these ground relays was the idea that ground faults are not as severe as phase faults and do not cause as much burning of the conductor by the arc nor lead to system instability. Accordingly it was permissible to take relatively long time delays for clearing. However, as one of the discussers said, they have found that faults will burn through two and three strands of 7-strand conductor and make it too weak mechanically to be trusted. Such burning would not be experienced with high speed clearing of ground faults.

As the author states, ground faults are in the majority running from 80 to 90 per cent of the total faults, and therefore it seems unreasonable that the cost of ground relaying should be expected to be only 10 to 20 per cent of the total relaying cost.

S. L. Goldsborough (Westinghouse Electric Corporation, Newark, N. J.): It is interesting to learn of W. A. Wolfe's experience with distance ground relays. His satisfactory results indicate that the applications were made carefully and that the limitations of distance ground relays were not exceeded.

Present day distance ground relays are subject to certain characteristics which have prevented wide application. These characteristics come about from the necessity of using a reactance distance element and the nature of ground faults. To measure distance on ground faults adequately, the reactance element is operated with a mixture of line current and residual current. In order to prevent response of the reactance element to load currents when the apparent reactance is small or negative, it is necessary to use a supervising impedance element. This impedance element must be set for the longest distance possible in order to allow the reactance element to function on faults involving high resistance, and yet not itself respond to load current. In other words the reach of the reactance element is limited by the reach of an impedance element. Where load currents are relatively high as compared to minimum fault current, this is a definite limitation.

A further limitation comes about from the action of the reactance element on 2-phase-to-ground faults. On this type of fault two relays receive fault current; one relay extends its reach and the other one restricts its reach. Since it is not possible to predict safely which relay will restrict its reach, it is necessary to arrange the impedance fault detector so that if more than one impedance element operates all three relays are blocked from tripping. This feature gives rise to the second limitation because it often happens that a single-phase-to-ground fault causes fault current to flow on all three phases, thus operating all three impedance supervising elements which block tripping. Therefore the reach of the reactance element is limited not only by load current, but by actual fault current.

Evidently the foregoing limitations were not exceeded in the applications made by Wolfe, but in any general application of distance ground relays these characteristics must be borne in mind.

Ben C. Hicks (Shawinigan Water and Power Company, Montreal, Quebec, Canada): About 20 years ago the Shawinigan Water and Power Company started to use impedance relays for single wire-to-ground faults on transmission lines. For this purpose phase-to-ground voltage has been used with either phase current or neutral current, depending on which is likely to give best results as indicated by calculations. This method of protecting against ground faults was adopted as a practical expedient and has proved to be reasonably satisfactory. The limitations of the method are recognized and admitted, however.

About three years ago serious consideration was given to converting a number of existing ground protections to the compensated form discussed by W. A. Wolfe. Authority was obtained to proceed with the conversion on two lines but because of pressure of other work the change has not been carried out yet. We therefore are interested very much in the results obtained by Wolfe.

In his paper Wolfe states that the second term in equation 3 represents a voltage in phase with the fault current. Presumably this statement is intended merely as an approximation and I would like to ask if the cases of overreaching that occurred are considered to be caused by errors arising from the approximation.

W. A. Wolfe: As Eric T. B. Gross suggests the use of two distance relays to do the work of six certainly would be attractive, especially on less important lines. It is difficult to see how the number of relays could be reduced without some increase in tripping time since the auxiliary switching relays for current and voltage first would have to recognize the fault and establish the proper connections to the distance elements before they could act to trip the breaker.

W. H. Cordray's discussion was welcome as coming from one of the manufacturers of distance ground relays, and it was gratifying to know that he agreed with the statements in the paper. On our system the use of high speed distance relays has resulted in very noticeable decrease in conductor and insulator damage from power arcs.

It is unfortunate that discussions were not received from some of the other users of distance ground relays, of which there are a few if not many. A letter from one such user indicates that their first three years experience with five sets of *HX* distance ground relays resulted in only 50 per cent correct operations. The author cannot help but believe that there was something wrong with the application or installation in this case.

S. L. Goldsborough presents an interesting discussion on the limitations of distance ground relays, most of which were mentioned in the paper. We do not consider the use of a reactance element instead of an impedance element a disadvantage except for the added complication in testing procedure when the phase relays are impedance and the ground relays reactance. When both phase and ground relays are the same, there is no disadvantage in this respect. Distance ground relays should not be applied in regions were high ground resistances prevail. In our territory we would expect no difficulty even on lines without overhead ground wires. On the type of relays having nondirectional impedance starting units, the danger of blocking tripping by fault currents flowing in the unfaulted phases is always a danger and must be checked by careful calculations before the relays can be applied safely. On the type of relays in which the impedance starting unit has an inherent directional characteristic, this danger does not exist as far as we know.

Ben C. Hicks asks a question in regard to equation 3. The second term is an approximation but, for all practical purposes, a close approximation. This term has the effect of pure resistance if $(I_{aGF} + I_{aHF})$ is in phase with (I_{relay}). In the usual case, all the impedances involved in equation 2 will have nearly the same phase angle unless the line resistances are very large. Also, since I_{aGF}, I_{0GF}, and I_{0GH} are usually nearly in phase for this type of fault, this will be nearly true. Probably variations in this term and the difficulty in determining exactly the amount of zero phase sequence compensation to apply, as expressed in the second term of equation 2, were the cause of the overreaching which we experienced. As stated in the paper, however, it was only necessary to decrease the distance setting of the relay to 85 per cent of the line length to prevent overreaching.

An error appears in equation 2 as published in AIEE Transactions. The second term should be $I_{0GF}\frac{Z_{0GH}-Z_{1GH}}{Z_{1GH}}$ instead of $I_{0GF}\frac{Z_{0GH}+Z_{1GH}}{Z_{1GH}}$.

Improvements in Performance of Hydroelectric Generating Units on the 2,000,000-Horsepower Saguenay System

Discussion and author's closure of paper 46-144 by F. L. Lawton, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 900–10.

J. F. Roberts (nonmember; Allis-Chalmers Manufacturing Company, Milwaukee, Wis.): F. L. Lawton has presented a very interesting and instructive paper. The combined Isle Maligne and Shipshaw plants form one of the largest concentrations of hydroelectric power at the present time. The speed of construction on the final Shipshaw plant, which placed 12 100,000-horsepower generating units into service, was a record as regards construction time.

The remarkably high efficiency of 94.4 and 94.5 on two of the eight Allis-Chalmers turbines in Shipshaw 2 plant we attribute not only to the smoothness of the welded plate-steel spiral casings, but also to the excellent water passages which were provided by the construction forces of the Aluminum Company. The penstock connection to the turbine comes in horizontally without any bends or elbows to cause disturbances. This is a powerful factor in increasing the efficiency of the turbines. The elbow draft tubes have a vertical length below the runners somewhat greater than is usual practice. This was possible because the rock foundations were somewhat lower than previously anticipated, and they were able to take advantage of this and build a more efficient type of draft tube without appreciable additional expense. These factors, we believe, contributed toward the high efficiency obtained.

Lawton's data showing the amount of metal removed from the various runners at the Isle Maligne plant also form a very interesting contribution. The amount of cast steel lost per year per runner runs from one-third to one-fifth of that lost by the cast-iron runners. This confirms other data and proves conclusively the superior performance of cast steel over cast iron.

It generally is admitted now that runners having the physical dimensions of Isle Maligne runners and operating under heads of 110 to 120 feet should not be made of cast iron. During the past few years a considerable number of large cast-iron runners have had to be replaced principally because of cracks that have developed where the blades join the upper crown. Cast iron lacks the ductility and toughness to resist the repeated bending action which occurs in runners of this size. During the original construction of the Isle Maligne plant it was necessary to economize wherever possible, and the fact that these cast-iron runners might have a shorter life than more expensive cast-steel runners was discussed openly. The maintenance men at the Isle Maligne plant have done a wonderful job in being able to keep these cast-iron runners in operation, as several of these runners have cracked near the crown, and if they had not been repaired properly these cracks shortly would have extended clear through the blades, with the total loss of the runner.

With our present knowledge of the shortcomings of cast iron, we doubt if any manufacturer or builder of large hydroelectric plants would consider installing a cast-iron runner at the present time. Lawton is to be complimented on an excellent paper which contains very valuable and interesting data.

H. R. Sills (Canadian General Electric Company, Peterborough, Ontario, Canada): F. L. Lawton has given some interesting data with regard to loss of capacity resulting from fouling of the machines. The curves shown in Figure 10 of the advanced copy do not conform in shape with loss curves or normal heating curves. In such heating curves, the temperature rise usually varies as the square of the load above a certain minimum determined by the no-load temperatures. The curves in this chart seem to vary as the square root of load, and if extrapolated to zero load would have a negative temperature rise.

It is apparent that there must have been error in making up this chart, which is to be regretted. There is need of published data on the effects of dirt in machine cooling passages.

It seems inevitable that there will be a significant proportion of machines built where the service conditions do not justify the installation of an enclosed ventilating circuit with air coolers, and in these the problem of maintaining the cooling efficiency with contaminated air still will be a serious one.

In respect to the machine in station 2 mentioned in the third last paragraph of the paper from which the large amount of dirt was taken from the inside of the rotor, it should be said that this was not the normal type of dirt which accumulates in service but consisted of cement dust and lint picked up during the construction of the power house. At this time this machine was arranged for open ventilation to provide a modicum of heat before there were any end sections in the power house.

The machines in station 2 had self-lubricated bearings and therefore were not plagued with oil vapor leakage, which almost invariably occurs on the older machines with an oil-circulating system. It is probable too that because of the prevailing conditions, the normal quota of honey bees and such viscous insects did not get into the machine to plug the ducts. Anyone who has driven far on a summer night will appreciate the adhesive qualities of certain insects.

Thus, while the cooling passages of the machine in station 2 were surprisingly clean (considering conditions), and the cooling performance was unimpaired, it hardly would be correct to say that it was a fair test. The general supersedure of the oil-circulating systems in machines by self-oiling systems will go a long way toward mitigating the cleaning problem. If the type of ventilation used at station 2 is as successful in separating out flies as it was dust and lint, some further improvement may be expected by this means.

R. A. Hopkins and **H. J. Petersen** (nonmember; Tennessee Valley Authority, Knoxville, Tenn.): It is noted that the original cast-iron runners at Isle Maligne were repaired by electric welding. It will be interesting to know what preparations were necessary and what methods were used in welding this cast iron. If stainless-steel rod is used for repairing cavitated areas, either on the original cast-iron runners or the later cast-steel runners, it will be helpful to know the analysis of the rod.

It appears that the specific speeds of the Isle Maligne and the Shipshaw 1 units bear about the same relation to the experience curve of limiting specific speeds, yet the Isle Maligne units cavitated much more than the Shipshaw 1 units. The author may have an explanation for this.

It is interesting to note that experience with open ventilation of the generators at Isle Maligne and at Shipshaw 1 indicated the use of closed ventilation at Shipshaw 2. As a parallel case, the eight generators of the original Muscle Shoals plant, built by the Army in 1925 and now operated by the Tennessee Valley Authority, have open ventilation. When a rotor of one of these machines is removed for other reasons, the stator is cleaned by air blast and scraping of the air passages, much as described by the author. A canvas hood is placed over the stator and exhausted upstream by a large vent fan to avoid contaminating other machines.

The 43 new units installed by the Tennessee Valley Authority at this and other plants have closed ventilation. Electric heaters are provided in the generator rooms to avoid opening the generators even during cold weather. Cleaning of these closed machines is required only occasionally and is done by a vacuum process at almost negligible expense as compared with the cleaning of open machines.

E. B. Strowger (nonmember; Buffalo Niagara Electric Corporation, Buffalo, N. Y.): The author has pointed out a number of improvements which have been made in the hydroelectric generating units of the Saguenay system, including both electric and hydraulic equipment. The writer is interested particularly in the turbine efficiency curves shown in Figure 16 of the paper, which shows the efficiencies of five large Allis-Chalmers turbines considered as outstanding water wheels in North America. The writer has participated in the Gibson tests of these units and wishes to make the following remarks in connection therewith.

The office of N. R. Gibson has made several hundred turbine tests throughout the world and based upon the results obtained, it seems that besides the consideration of runner design there are two factors which are believed to have an important bearing on the efficiencies of hydraulic turbines. The evidence is not always consistent but there is enough of it to attract attention, and the exceptions possibly may be explained in other ways or may be only sufficient to prove the rule. These two factors are

1. Constant diameter bends in the penstock near the entrance to the turbine.
2. Valves of a type such as butterfly valves which tend to disturb the flow of the water as it approaches the turbines.

These adverse conditions are mitigated by making a bend in the penstock, converging so as to accelerate the flow around the bend and by using a valve like a gate valve which, when open, leaves the waterway unobstructed.

Referring to Figure 16 of the paper, the Shipshaw turbine unit 9, has a straight water passage leading to it with no lower penstock bends and no valve. It has a Y at some distance upstream, but the flow in the Y is accelerated. The Niagara unit 21 has no bend near the turbine and the penstock valve is a Johnson valve with concentric water passages and accelerating flow. Unit *L*7 at Grand Coulee has two lower penstock bends, one vertical and one horizontal, both of which are close to the turbine although the flow is accelerated. The Osage unit involves the common design of penstock with reversed curve through the dam, and Boulder unit *A*5 has a butterfly valve located immediately ahead of the turbine.

There also should be considered in an analysis of this kind the factors of head and quantity of flow. Based upon many years of experience in testing work, a head of about 200 feet, associated with a turbine designed to discharge a relatively large quantity of water, seems to produce ideal hydraulic conditions for maximum turbine efficiency. Generally, as indicated by Figure 16 of the paper, as the head departs in either direction from this value, maximum turbine efficiency seems to decrease.

F. L. Lawton: The discussion by J. F. Roberts adds to the author's presentation of the essential details of the Shipshaw 2 water wheels by stressing those factors contributing to the high efficiencies obtained. The comments on the relative merits of cast iron and steel for large moderate-head runners, both as regards resistance to pitting and resistance to cracking of the vanes near the crown plate, set forth the identity of viewpoint of manufacturers and most operators. The greater ability of cast-steel runners to resist pitting and especially cracking, and ease of effecting satisfactory repairs, are resulting in most new runners being cast steel.

H. R. Sills's remarks on the effect of dirt on the cooling system and performance of hydroelectric generators are welcomed. It is appreciated that the curves of Figure 10 appearing in the advance copy of the paper are somewhat in error. A re-examination of the original data shows these should be redrawn as in Figure 10 of the published paper. While the change alters the quantitative values somewhat, the basic assertion of Sills's discussion is supported. It should be noted that the original tests were designed to show relative temperature rises under similar conditions rather than absolute values.

R. A. Hopkins has inquired as to necessary preparations and technique used in welding the original cast-iron runners at Isle Maligne.

In affecting repairs to pitted areas, initial practice was as follows:

1. The pitted or spongy metal was chipped out to sound cast iron, using pneumatic chipping hammers.
2. The chipped area was drilled 3/4 to 1 inch deep and tapped for 5/8-inch by 1-inch cap screws on approximately 4-inch centers. These cap screws were tightened, cut partly through under the head, and tightened again until the heads broke off.
3. A priming coat of low-carbon steel was laid down by the electric-arc process on the chipped area and carefully peened as applied. That is, care was taken to limit the area welded prior to peening, which was thorough rather than heavy.
4. The cavity created by chipping out the spongy metal was built up by depositing a low-carbon steel closely to the original contours. Each layer was peened prior to deposition of the succeeding layer.
5. A finish coat of high-carbon steel was laid down in two layers, carefully peened, and ground to the original contours as closely as practical.

Early experiments with a stainless-steel finish coat on areas subject to pitting were not too encouraging, but developments in welding rods and in application technique have progressed to the point where current practice in welding cast-iron runners is

1. The pitted or spongy metal is chipped out to sound cast iron, using pneumatic chipping hammers.
2. The chipped-out area is drilled, tapped, and studded as previously described, latest practice tending to somewhat less studding.
3. A priming coat of low-carbon steel is laid down on the chipped area and peened. Somewhat better bond is secured with currently-used rods than was formerly possible.
4. The cavity resulting from the chipping operation is built up by deposition of successive layers of low-carbon steel laid down by the electric arc, successive layers being peened.
5. A finish coat consisting of two layers of 18–8 stainless steel is laid down with the electric arc, each layer peened, and the finish layer ground to contour. The average analysis of the deposited metal in the finish coat is 0.07 for carbon, 19.70 for chrome, 9.50 for nickel, and 0.80 for columbium. Reverse-polarity rods with an extruded coating are used.

Pitted areas on cast-steel runners have been repaired in much the same manner as for the cast-iron runners, but without studding and using a reverse-polarity extruded-coating mild-steel electrode.

Cracking in the vanes of the Isle Maligne cast-iron runners followed quite a consistent pattern. Cracks almost invariably started at the discharge edge of the vanes or buckets and ran outwards towards the inlet edges, sometimes as far as 36 inches. Beginning at a point 2 to 4 inches down from the crown plate, the cracks would terminate in the crown plate.

The method of repair originally was as follows:

1. A V-shaped cut was chipped out along the crack on the face of the vane, using pneumatic chipping hammers.
2. Holes were drilled into the two faces of the V-shaped cut, on $1^{3}/_{4}$- to 2-inch centers, and tapped for 5/8- or 3/4-inch diameter steel cap screws. These were screwed tight, cut partly through, and the heads broken off by tightening still further.
3. A low-carbon steel electrode was used for the deposition of a priming coat on the surfaces of the V-shaped cut. The priming coat was peened lightly and the space gradually built up somewhat beyond the original contour using a medium-carbon electrode, peening successive layers. Initially a coated English rod was used, but later experience indicated comparable results could be obtained with Canadian rods at lower cost.
4. The rough surface of the weld was ground off to contour.

This technique has been employed successfully for replacement of an entire vane in one of the 128-inch-diameter 67-specific-speed Isle Maligne runners. In this case a double V-shaped cut (from both faces) was employed. The replacement vane was cast steel. The repaired runner has given 18 years service without any trouble insofar as the new vane is concerned.

However, it is extremely difficult to establish a bond between the parent cast-iron metal and the weld deposit, which will transmit permanently the heavy bending stresses from the shroud ring and vanes to the crown plate. Consequently, our present practice is to bridge the crack by steel plates secured to the crown plate or upper portion of the vane, and to the vane below the crack by heavy studs with countersunk heads welded in place. The steel plates also are welded to the cast-iron surfaces around their entire periphery.

Some cracking of cast-steel runner vanes, similar to that described for the cast-iron runners, has been encountered at Isle Maligne. Originally the cracks were chipped out, leaving a V-shaped cut. This was studded in similar fashion to the cast-iron runners, and the space gradually built up with successive weld layers, using a reverse-polarity extruded-coated mild-steel rod corresponding to American Welding Society's classification *E*6010. However, such repairs were not too permanent, "lifting" occurring at the bond, so that present practice is to keep the working area warm at all times, depositing weld metal in layers, and carefully peening each layer.

Hopkins' query as to the reason for the greater pitting experienced with the Isle Maligne runners as compared with the Shipshaw 1 runners is interesting. Both designs bear roughly the same relation to the experience curve of limiting specific speeds. Both have about the same relative position to the experience curve for limiting values of the cavitation factor, sigma. However, the Isle Maligne units were designed with a specific speed as high as was considered practical then for high initial economy, whereas the Shipshaw 1 units were designed for operating economy. The higher head at Shipshaw 1 is conducive to better efficiency and, as a corollary, less pitting. Operating conditions influence the rate of pitting, as indicated in the paper, especially when units are operated for long periods at gate openings over best gate. Another significant factor is believed to reside in the shape of the water passages through the runner, but its relative importance is unknown.

The use of the expression "pitted" to denote areas damaged as a result of cavitation is desirable in the author's opinion. "Cavitation" properly describes the formation, in the mass of water moving through the water wheel passages, of cavities or voids filled with air and water vapor.

The reference by Hopkins to the Tennessee Valley Authority practice in ventilating and cleaning generators is most useful, especially the comment on use of electric heaters in the generator rooms to avoid opening the generator ventilating systems during cold weather.

E. B. Strowger's discussion is particularly interesting in view of the extensive background of water wheel efficiency testing behind it. There is little doubt that smooth water passages from forebay to tailrace, with properly-designed bends, are necessary for attainment of high efficiencies. The description of the water passages leading to the water wheels on which Figure 16 of the paper is based is helpful to a full understanding of the relative efficiencies. The significance of the head and quantity of water on efficiency of large water wheels, as noted by Strowger, should not be overlooked.

Comparisons of Railway Motive Power for Operations in the Pacific Northwest

Discussion and author's closure of paper 46-186 by T. M. C. Martin, presented at the AIEE Pacific Coast meeting, Seattle, Wash., August 27–30, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1054–65.

Walter S. Gordon, Jr. (consulting electrical engineer, Tacoma, Wash.): T. M. C. Martin's paper is a well-prepared brief on a difficult subject. Greater detail would have obscured the picture he has presented so clearly. Such a simplified approach is required emphatically to gain the hearing and serious consideration of railroad officials skilled in railroading with many years of experience in the utilization of steam motive power. Practical operating officials on railroads, not electrical engineers, must be convinced of the economic and operating advantages of electrical operation.

During past years many electrical engineers and others interested in electric traction appeared to be more interested in presenting the merits of their one particular system of electrification than they were in presenting the over-all advantages of all systems of electrification. It is not to be wondered that, as a result of this technical conflict between electrical ideologies, top railroad personnel charged with making decisions regarding the installation of electric traction hesitated and waited for the smoke of battle to clear. Under similar conflicting circumstances our American industry most certainly would not have become as highly electrified as it is today. It is indeed fortunate that each individual plant or mill did not have to arrive independently at the solution of such problems as the correct transmission and distribution voltage, whether to use direct or alternating current, and if alternating current, the phase and frequency. And then face the delays and expense of securing motors, transformers, control, and switchgear specially designed and constructed for each individual installation!

Most railway electrifications have been conceived brilliantly and executed well in a daring and competent manner. The principals deserve the highest praise for their able work, which required much technical pioneering to overcome many technical difficulties during design, construction, and the initial phases of operation. But too many other electrification installations, despite ample economic justification, never reached or passed the planning stage.

In the past, much emphasis was placed on the details of selecting an electrification system that was technically correct for a specific section of railroad under certain assumed traffic and operating conditions. Later and unforeseen changes in traffic and operating methods voided many of the assumptions considered vitally important in the original selection. For example, strong argument now could be advanced for installing alternating current on what is now the longest d-c electrified American railroad, and likewise for installing direct current on what is now the longest a-c electrified American railroad! However, the superior operating and economic performance of both these outstanding electrifications cannot be denied now. Fifty years of electric traction have proved its inherent merits.

As the representative electric locomotive for this study, Martin's selection of the motor-generator type appears most reasonable and should meet with a minimum of dissent. Considering the service requirements on Pacific Northwest railroads, this type of electric locomotive has the strong advantages of being usable interchangeably in both freight and passenger service, of 20 years successful operating experience in this area, of making available at the rail the full horsepower capacity over almost the entire range of operating speeds (from 17 miles per hour on mountain grades to 60 miles per hour on light ascending grades) and of trainweight regenerative braking over the same wide range in operating speeds. However, this does not imply that the motor-generator electric locomotive should be selected for all future electrifications, wherever situated. For other regions with different physical, operating, and traffic conditions, it is possible that one of the other types of electric locomotives such as the a-c series motor or the phase-converter would be used; it is also possible that one of the many other arrangements of steam locomotives likewise might be selected as the representative type for that form of motive power. However, at present the motor-generator locomotive comes closest to being a universal a-c electric locomotive. It is significant that the motor-generator locomotive is now satisfactorily in use on an eastern railroad which has high-speed passenger and freight traffic on level and light-gradient routes, and which for 37 years has operated a-c series motor locomotives; also that four motor-generator locomotives have been ordered recently by a southeastern coal-carrying road which has a heavy-tonnage freight traffic on steep mountain grades, and which for more than 20 years successfully operated a-c phase-converter electric locomotives. Thus the author's selection appears well documented by recent practice elsewhere in this country as well as by the 20 years of operating success achieved on an electrified railroad in the Pacific Northwest.

The author's selected type of a-c power distribution system already has been installed on by far the most miles of track and route in this country. Operating today from the 12-kv 25-cycle single-phase a-c catenary system in the east, south, and west are electric locomotives which have traction motors of three widely different types and characteristics. These types are the single-phase a-c series commutator motor, the 3-phase dual speed a-c motor supplied from a locomotive-borne phase-converter, and the d-c series commutator motor supplied from a locomotive-borne synchronous motor-generator. Such a distribution system therefore provides a maximum of flexibility for selecting a type of motive power to suit any railroad's profiles, traffic, or operating conditions and methods. This system has a large capacity, although the necessity of supplying 15,000 kw to any point on the 12-kv trolley, as mentioned in the paper, might be questioned as somewhat high. Power in the order of 7,500 kw at present, and of 10,000 kw for the future, appears to be sufficient.

Martin's observations regarding a choice of frequency were most interesting. As he suggested, American practice would be better today if 15 or 20 cycles, instead of 25 cycles, had been established long ago. It is probably impracticable to change now to one of these lower and more desirable frequencies despite the many recognized advantages. However, to the discusser it does appear reasonable and practicable now to adopt 24 cycles as a national standard traction frequency for use on all future electrifications; this would mean a deviation of only four per cent from the present accepted standard. As mentioned in the paper, the use of 24 instead of 25 cycles would result in considerable savings in the costs of frequency changers; the maximum speed which could be used for these motor-generator sets then becomes 720 rpm instead of the present 300 rpm in converting from 60 cycles. Locomotives and equipment designed for 24 cycles would operate equally well on 25 cycles, and it is believed that no serious difficulties would be encountered in actually operating existing 25-cycle locomotives from a 24-cycle trolley system. This proposed 4-per-cent deviation in frequency is less than half the present 10-per-cent deviation in frequency permitted, with satisfactory performance guaranteed, by present commercial standards for industrial motors. In the interest of economy on future electrifications, this slight change in traction frequency should be adopted now by all manufacturers of traction equipment.

The author has leaned backward to attain objectivity and insure equity to steam and Diesel motive power. Since railroad operating officials that are now steam-minded or already favorably impressed by the many advantages of Diesel-electric traction have to be convinced that in many instances electrification economically will solve their operating problems, such an attitude is reasonable in such a study. It is better to claim too little than too much. However, it should be noted that in the comparative cost of facilities (Table XI of the paper) and the operating and fixed charges (Table XII) for steam operation, the costs of fuel and water stations and similar facilities have not been included; and in the timetable comparisons of maximum performance (Table VII) no allowances were made for inevitable delays caused by stops for water and fuel when steam locomotives are used. Locomotive depreciation rates appear to favor steam locomotives unduly. For coal, a value of $2.00 per ton at the locomotive appears quite low as this price must include the cost of the fuel at the mine, the cost of transportation from mine to fueling stations, and the cost of storage, handling, loss, and waste. For example, the cost of coal purchased by a large eastern seaboard utility was an average of $5.89 per ton in 1944 and $6.12 per ton in 1945.

A value of four mills per kilowatt-hour for hydro-generated electric energy at the frequency-converter terminals, mentioned in Tables XII and XIII of the paper, may seem unrealistic to many power and railroad personnel familiar with the much higher energy rates that railroads generally are paying now. In some instances these present high rates are relatively far out of line with rates for industrial energy. This seems especially true in the case of electrifications where the railroad's own commercial frequency transmission and transformer facilities are made available to the power suppliers for use in selling commercial power and light to other customers. The eastern utility mentioned previously generates about 80 per cent of its total electric production from coal, which in 1945 cost $6.12 per ton, and receives an average of nine mills per kilowatt-hour for industrial sales and less than eight mills per kilowatt-hour for energy sold to electrified railroads and street railways. This railway load, which constitutes 20 per cent of the utility's total energy sales, is considered profitable. A western utility, with lower-cost hydro-generated power, receives an average of six mills per kilowatt-hour for industrial sales and much more than that for railway sales, although on a relative basis a railway rate of about five mills would be in order. This is approximately the rate quoted by a western utility more than 30 years ago, and since that time, industrial rates have been halved at least. The 4-mill rate mentioned by the author, which appears extremely low now, may prove to be an attractive source of income for some utilities. In any event, this 4-mill rate for railway electrification energy soon may be an established fact here in the Pacific Northwest.

Finally, as a matter of national policy, utilization of our hydro-generated electricity instead of consuming our apparently dwindling liquid fuel resources soon may be recognized generally as desirable and necessary. At the present time such utilization should be encouraged at least wherever it is economically sound and technically feasible. Oil replaced at home by water power should be much more valuable than oil rights in distant foreign lands. Therefore the lower rate of return on the incremental investment required for electrical as compared with Diesel operation (Figure 7 of the paper) could be compensated justifiably by making available to railroad managements long term loans repayable upon the basis of extent of use as measured by energy consumed (which would vary with traffic or ability to pay) instead of upon the basis of a fixed annual sum repayable without regard to the extent of use. A practicable basis for repayment is vitally important to railroads if they are to be able to afford the higher initial cost of electrification. When traffic is heaviest the railroads need electrification and have the money to pay for it, but they are usually then too busy to install it; also, labor and materials are then sometimes difficult to secure because of a national crisis, as during a war. During depressions, traffic is lower, the economic need for electrification decreases, and net revenues are low, so that private funds become difficult to secure. Therefore electrification should be expected to pay off the investment on the basis of the average traffic over a reasonable period of years, at least a decade or longer.

The discusser in 1932[1] wrote that

> "European railways are apparently well satisfied that the operating economies, increased speed, cleanliness, increased track capacity, and use of natural resources such as water power or low grades of coal, are advantages that justify further extension of railway electrification even during a severe economic crisis. The execution of such projects during times of depression has been of material assistance in alleviating unemployment in several of the countries. Perhaps America could profitably follow such an example."

Is it now too much to expect that the more progressive of our railroad managements, the electric and railroad equipment manufacturers, the interested power utilities, and government transportation agencies will at last get together and co-operate on a sound long-range program of expediting the electrification of all railroad lines which will justify the investment? Specific and detailed analyses then will reveal the railroad lines which should be electrified.

REFERENCE

1. A REVIEW OF EUROPEAN RAILWAY ELECTRIFICATION, W. S. Gordon, Jr., E. Jaboolian. *Electrical Engineering*, April 1932, page 252.

H. F. Brown, (New York New Haven and Hartford Railroad, New Haven, Conn.): T. M. C. Martin, in his paper, has presented a very clear comparison of the three outstanding types of motive power available today for trunk-line railroad operation. The railroads of the Pacific Northwest, with their heavy mountain grades, present a problem for which the electric locomotive is one of the proper and logical economic solutions.

It is the opinion of this commentator that Martin's analysis and conclusions as to the type of distribution of the electric power supply, that is, 12 kv, 25 cycles, single-phase alternating current is sound, and that the choice of the motor-generator type of locomotive with d-c motors is also sound because of the necessity for lower speeds with maximum tractive effort on long grades, the possibility of regeneration especially for braking, and the possibility of power-factor correction with the synchronous motor-generator sets.

The electrification of these railroads presents an investment problem that the purchase of self-contained types of motive power do not, namely, the possibility of financing the necessary capital investment through equipment trusts. However, it is not outside the range of possibilities that where electrification can be applied with superior economies in a territory where cheap electric power from natural resources can be secured, some similar means can be developed for financing the power distribution system.

The power engineer cannot but be impressed by the various power conversion problems which are presented in the electrification of these railroads. First, the generation at 3 phase, 60 cycles alternating current; then the conversion to 25 cycles, single-phase alternating current for the railroad contact system; and finally the conversion on the locomotive to direct current for the most favorable motor design. If the railroad load is a legitimate field for the power development companies and administrations, then their plans for the development of the natural resources should include provisions for the direct generation of 25-cycle power for such purposes, thereby saving additional investment in conversion equipment and the attendant losses in power. Twenty-five-cycle power can be utilized for certain commercial consumption just as economically and efficiently as 60-cycle power; for example, aluminum production, and it seems rather shortsighted to put all the investment into 60-cycle generation if the electrification of railroads is a sizable part of the ultimate load the power development hopes to assume.

Charles Kerr, Jr. (nonmember; Westinghouse Electric Corporation, East Pittsburgh, Pa.): T. M. C. Martin's paper, while somewhat provocative, presents a very intelligent discussion of railroad electrification. Few with operating experience on modern electrified railroads will dispute the claim that electric operation offers many outstanding advantages over any other known type of motive power. The only deterrent to extensive electrification has been first cost, and if the differential in this item between electrification and other types could be reduced materially or eliminated, the field of railroad electrification would be virtually unlimited. Reduction in capital investment is the paramount assignment for all parties interested in the cause of railroad electrification.

Without desiring in any way to initiate a "battle of the systems," Martin's selection of electric system and locomotives for universal application is open to debate. The 25-cycle a-c system, employing motor-generator locomotives, was introduced by Westinghouse, and we would be the last to deny its many virtues or its complete adequacy under the operating conditions encountered in the Pacific Northwest. However, certain recent analyses of systems of electrification for quite similar applications to those anticipated by Martin have demonstrated that the a-c system with motor-generator locomotives well may involve the largest capital expenditure. The motor-generator locomotive was introduced in 1926, and many advances in the art have taken place since that time.

There are several systems of electrification which, when properly engineered, would handle adequately traffic in the Pacific

Northwest. Therefore, no system is the only answer from the standpoint of adequacy to handle the service. Furthermore, in general, the operating expenses between the various systems differ only to a minor degree. Thus, the ultimate selection can be made essentially in favor of that system which requires the minimum capital investment when properly engineered to meet the over-all service conditions of this territory.

It may be of interest to point out that the 25-cycle a-c system, employing commutator motor locomotives, has been used most extensively of all types. This system has handled the most dense traffic in America for many years. Its operating flexibility never has been matched. The same type of locomotive, operating from a 15-cycle system, could provide this same flexibility, and, in addition, meet the heavy-grade requirements of the Western roads in an equally satisfactory manner. There is also considerable evidence that this system would require a definitely lower first cost. With future technical developments not too far removed, other systems might fulfill well western operating requirements with advantage.

It is my belief that Martin's aim of extensive railroad electrification in areas of assured low cost power can be furthered best by selecting that system which meets the requirements of this territory with the absolute minimum of initial investment. I question seriously whether the system proposed in his paper accomplishes this purpose.

Electrification projects involve the interested railroads, power companies, and manufacturers. It might be well for these groups to find some means of reaching a common standard which would provide the most economical over-all system for the Pacific Northwest, and such action might prove to be a considerable factor in extending electrification in this region. The land transportation committee of the AIEE might initiate this co-operative effort.

T. M. C. Martin: I am interested very much in Charles Kerr's discussion. I agree completely with his observation that the best system is the one that provides the lowest over-all costs—lowest for the railroad and lowest for the power supplier.

I am unable to perceive that 15 cycles offers very startling economies at this late date, but be that as it may, I am convinced that the use of a-c series motors in this part of the country is not only prejudicial to the best interests of power suppliers but it offers largely illusory benefits to railroads. The chief claim presented for their use by Kerr seems to be lower first cost. Suppose we examine locomotive costs on the basis of their cost per pound and their cost in terms of dollars invested per thousand pounds of continuous tractive effort. All we have to go on are facts in the form of recent prices for representative locomotives taken from each class.

From Table I, and keeping in mind the fact that continuous tractive effort is the only real measure of performance that matters to western railroad managements, it is very difficult to follow Kerr's lower first cost contentions on behalf of a-c commutator motor locomotives.

Table I

	15-Cycle 24-Kv Trolley A-C Commutator Locomotive	25-Cycle 12-Kv A-C Motor-Generator Locomotive
Manufacturer's recent asking price	$475,000	$500,000
Total weight of locomotive, pounds	600,000	720,000
Price per pound	$0.79	$0.69
Maximum continuous tractive effort in pounds	99,000	119,000
Price per 1,000 pounds of maximum continuous tractive effort	$4,798	$4,202

If this is true it is even more difficult to perceive a justification for their use in terms of annual operating costs. Most authorities agree that their maintenance will be more costly, and there can be little doubt but their power supply will be more expensive for at least two very simple reasons:

1. Their single-phase power factor will be poor, probably no better than 80.0 percent at the best.

2. Their ability to assume short-time overloads and their relatively higher minimum speeds for maximum power output cannot fail to produce a lower working electric load factor. This latter point is an especially difficult one to sidestep because western electrification is at best a fairly low load factor proposition.

Performance Criteria for Current-Limiting Power Fuses—I, II

Discussions and authors' closure of papers 46-170 and 46-171 by C. L. Schuck and E. W. Boehne, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, pages 1028–45.

E. A. Williams, Jr. (General Electric Company, Philadelphia, Pa.): The testing of a line of fuses to establish interrupting ability over a wide range of currents has been, in the past, a long and arduous task because of the effect of circuit variables upon interrupting performance, and because of the large number of ratings involved. To accomplish the desired results with any degree of certainty by cut and try methods is almost impossible, as the operation of most fuses is quite critical under certain combinations of circuit conditions.

The papers by E. W. Boehne and C. L. Schuck recognize and discuss the factors controlling fuse performance, establish a definite mathematical method of predetermining areas of critical performance, and present adequate test data to substantiate the accuracy of their theoretical approach to the problem.

The value of the papers is threefold:

1. The basic data presented contribute to a better understanding of the phenomenon of circuit interruption in general.

2. The methods of procedure described provide the fuse designer with a useful tool to produce better fuses at lower cost.

3. The user may apply fuses with greater assurance of reliable performance under any fault condition within the rating of the fuse.

H. H. Fahnoe (Westinghouse Electric Corporation, East Pittsburgh, Pa.): C. L. Schuck and E. W. Boehne in their papers have given the fuse designer a valuable review of the many factors that must be considered in developing a current-limiting type of fuse. The data presented are also of interest to the engineer making application of these fuses at important locations on central station and high-power industrial systems because they provide assurance that this type of fuse is capable of interrupting all magnitudes of short-circuit currents experienced on his system.

Part I outlines certain laboratory test requirements and procedures. Surely, the broader aspects of these criteria were recognized during the development of the fuse described in reference 2 of part I by Prince and Williams. However, Prince and Williams did not include any information on test procedures in their paper. In reference 10 of part I, Rawlins and Fahnoe described in detail the test methods followed in the development of a current-limiting fuse. It is interesting to note that part I of these papers substantiates the test procedures and the need for extensive high-power laboratory testing as set forth by Rawlins and Fahnoe.

It has been recognized for many years that the high-power testing laboratory offers the only completely satisfactory means of proving the interrupting ratings of power fuses. In the early days, the fuse designer was forced to rely on the incomplete tests which he could obtain on commercial power systems. With the advent of the high-power laboratory, where all factors affecting the test circuit can be controlled and measured, the development of the highly efficient modern circuit interrupter became possible.

This is particularly true in the case of the current-limiting fuse. In a faulted circuit containing this type of fuse, the fuse is the controlling element in the circuit because of the high arc voltage associated with its operation. Consequently, the effect of parallel loading, recovery voltage rate, power factor, and the like are of less importance than with other types of circuit interrupters. More particularly, as stated by Rawlins and Fahnoe, it is necessary with current-limiting fuses to test "all ratings of fuses...at rated voltage with all conditions of asymmetry on currents of varying magnitudes from minimum melting current up to the interrupting rating of the fuse." Thus the tremendous number of tests required to explore the performance of a line of fuses adequately practically eliminates the possibility of field testing. Because of the completeness of the laboratory tests, the need for field testing is obviated.

It is gratifying to note that the conclusions regarding the testing of current-limiting fuses stated in part I of these papers agree remarkably well with those set forth in the Rawlins and Fahnoe paper, and in no respect are the conclusions contradictory. Such agreement between independent observers helps immeasurably to verify conclusions, thereby establishing the fundamentals of the art.

J. W. Gibson (The General Electric Company, Ltd., Birmingham, England): The design of the current-limiting fuse has been

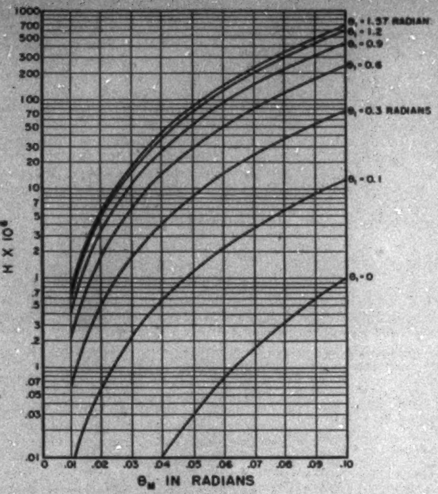

Figure 1. Heating function curves, supplementing Figure 8 of part II, which will be useful for fuses of low current ratings, and fuses of moderate current ratings under high fault current conditions

The H functions are computed from the relation

$$H = \frac{\Sigma i^2 t . \omega}{I^2}$$

The melting time in seconds is equal to the melting angle in radians divided by 2π times the circuit frequency. The curves are a plot of equation 22 of part II with minor corrections to align with equation 5

facilitated greatly by analytical and experimental investigations of its current-breaking performance during the past 10 or 15 years, and the present papers mark a considerable step forward in this direction. It is becoming possible to design the fuse for any required interrupting performance, largely avoiding the method of trial and error, and in addition to predict its discrimination performance when used in series with circuit breakers or with other fuses.

I agree with the authors that arc energy is the most generally useful criterion of the severity of a test. It should be remembered however that other factors may supervene to cause failure. For example, if the rate of rise of fault current is high, the excessive mechanical pressure set up may cause failure through bursting of the cartridge before the fuse has accepted a value of arc energy with which it can deal successfully if the rate of rise is less.

Figure 4 of part I illustrates well that the current-limiting fuse is more sensitive to changes in circuit voltage than to changes in short-circuit current; this property is the reverse of that found with most other interrupters.

The curves of Figure 5 of part II show a diminution of arc energy when the arc voltage falls below about half of the crest value of the applied voltage. I should be glad of the authors' view as to the possibility of a powder-filled fuse operating successfully in this zone; I should have expected that the fuse would be unable to interrupt either at the first, or at subsequent current zeros, resulting in failure.

Will the authors give their experimental value for their constant F_m (equation 7 of part II) for silver? In my experience, the value increases slightly with increase in the current density in the element at the initiation of arcing, because the melting process tends to proceed further before a break in the element is initiated. The curves of Figure 8 of part II, derived from equation 16, will render it easy to determine accurately the pre-arcing time for various starting angles. It would be useful, however, if the author would extend them downwards to cover the case of fuses of very low current rating, where the melting time is only a very few degrees and equation 22 applies.

The conclusions drawn in the papers regarding the effect on severity of fault angles and of available ("prospective" in British terminology) current line up well with findings in England. Thus, for certain British Standard specifications in preparation, it is proposed that breaking tests shall be made, not only at rated megavolt-amperes, but at a less value of megavolt-amperes near which the peak value of arc energy may be expected to occur, the criterion being that the prospective current shall be so chosen for the test that the current at the initiation of arcing is between 70 and 100 per cent of the peak symmetrical value of the prospective current. Further, certain stipulations probably will be made regarding fault angles. In a British Standard for high voltage fuses, a proposal is that arcing in current-limiting fuses shall commence on a rising voltage between 50 and 100 per cent of peak value. For lower voltage (up to 660 volts) fuses, with current ratings up to 1,200 amperes, a proposed stipulation relates to the time of current initiation, and not to arcing initiation; the circuit is to be closed on a rising voltage between 35 and 65 per cent of peak value. In the case of fuses of low current rating, this will result in arcing commencing on a rising voltage. For larger fuses, arcing will commence on a falling voltage, as it is believed from test experience that this is likely to result in the most severe condition. The reason is that a current zero (following a minor current loop) is reached shortly after arcing commences, and as a result the arc paths have not had sufficient time to deionize themselves. This factor, together with the relatively low current density in the arc (arising from the use of numerous elements in parallel) results in a second loop of current, with consequent increase in arc energy before arc extinction occurs. I should like to know whether the authors have experienced this phenomenon. The point also may be of importance in connection with fuses used in circuits, for example, furnace circuits having frequencies of the order of 1,000 cycles per second. My experience with such frequencies agrees with the analytical conclusions of the authors that increase in frequency results in reduced severity, but it is possible that above certain frequencies continued arcing may occur on account of high restriking rates at the zero pause. In other words, zero-pause phenomena, of a type not normally associated with current-limiting fuses, may assume importance. I should like to have the authors' views on this subject also.

Figure 10 of part II shows in an effective manner that arc energy diminishes as the fault current increases above a certain value. It would be useful if the authors would give similar curves for varying arc voltage because it appears certain that the characteristic just mentioned would not apply to the lower values of arc voltage.

E. W. Boehne and C. L. Schuck: The authors greatly appreciate the interest shown in these papers, particularly because those who have submitted discussions have been connected closely with power fuse development.

E. A. Williams, Jr., describes the vast number of fault conditions encompassed by a single number expressing the interrupting rating. The fuse designer, confronted as he is with the variables of the circuit on the one hand, and with the variables involved in the fuse design on the other, seeks a procedure that will give him maximum information with a minimum number of tests. His interest therefore centers in the critical circuit and design factors, which he perceives by the joint exercise of a subjective analytical approach and an objective correlation of test results.

In H. H. Fahnoe's discussion, he mentions certain references made to test procedure in earlier papers. The subject of fuse test severity is certainly not new, and, if there is interest in earlier contributions to our present knowledge, surely the work of Gibson,[7] Metcalf,[9] Grant[19], and others should be noted, together with that of Lohausen[8] and Lapple[5] (German), and Grillet[18] (French). (All references are in part I of the paper.) All of these predate the brief reference to test procedure made in the Rawlins and Fahnoe paper.[10] In that paper, the importance of causing melting to occur near voltage peak was reiterated,[7] but this factor must be considered jointly with the very important severity factor of the magnitude of current at melting. Figure 5 of part I shows the effect of fault starting angle on this factor. For fuses of low current ratings, it so happens that maximum current, voltage, and severity all obtain for a fault starting angle slightly before 90 degrees. For fuses of high current ratings, maximum melting currents are obtained if melting takes place at or even after voltage maximum. Thus, if the fault starting angle were chosen only to give maximum melting current, or to bring about melting at voltage peak, the interruption would be made easier, because arcing

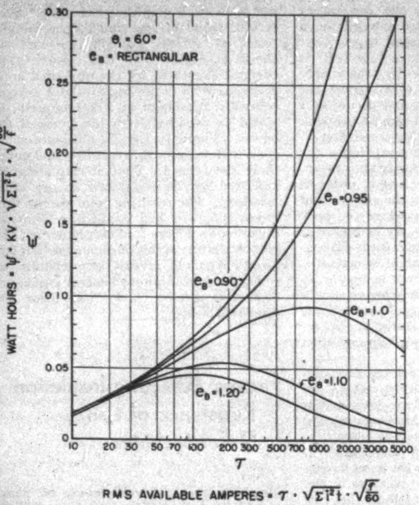

Figure 2. Arc energy characteristics resulting for magnitudes of rectangular arc voltages ranging from 90 to 120 per cent of the crest of the generated voltage

These relations are for zero power factor short circuits initiated 60 degrees following a voltage zero. They hold for any fuse rating, any frequency, and any voltage. These relations emphasize the importance of the arc voltage of the fuse, especially for high values of short-circuit current. Such characteristics are useful in the adequate testing of fuses

would take place during a time when generated voltage was decreasing (see Figure 6*B* of part I). Evidently a more accurate and quantitative statement was needed on this question, and these papers give this in the form that melting should occur in the region 70 to 80 degrees after voltage zero. On this basis, the fault starting angle, rather than being zero, should be related to the current rating of the fuse. For the 20,000-ampere circuit of Figure 5, melting takes place at approximately 75 degrees if the circuit is closed at the values given in the following table:

θ_1, Degrees	Fuse, Amperes
0	200
42	100
60	50
72	7

It follows from this table that fuses having current ratings in excess of 200 amperes on this circuit would find the most difficult conditions when θ_1 is negative, that is, under conditions of an initial minor loop, melting occurring well into the next loop of current. In some cases of high current fuses, fusion late in the initial half cycle will not be followed by interruption at the subsequent voltage zero, and an additional loop of current will result. This adds to the duty and requires special consideration, as Gibson points out.

The papers also include a consideration of each of several additional factors which are properly a part of the subject of power fuse performance because both designer and user must know their effects. Questions arise on the effect of frequency, power factor, polyphase versus single phase, and a-c versus d-c testing, and it was believed that a clear statement embodying our present knowledge on each was desirable.

Gibson has made several thought-provoking comments, which indicate the thoroughness with which these questions have been considered in England. It is significant that certain conclusions reached there on test severity, dictated by a careful analysis of test data, and which agree so closely with our analysis and experience, probably will be included in the British Standard specifications.

He inquires as to the constancy of the coefficient F_m in equation 7, part II. A progressive decrease in this coefficient with increasing element diameter has been found empirically. For a 200-ampere fuse (employing large-diameter elements), the value of the coefficient is 0.0188, that is, 12 per cent below its value of 0.021 for low current ratings employing small-diameter elements (the element metal being silver). The value 0.021 is also that calculated on the basis of a heat input sufficient to raise the temperature of the silver to its melting point, and to convert it to a liquid at that temperature, no heat being lost. The decrease in F_m suggests a mechanism initially resembling skin effect, with current concentrating in the outer portions of the element cross section. It is shown in the following text that such a nonuniform current density alone can shorten the melting time and make F_m appear smaller. There also may be a local shattering effect, as the current transfers to the lower-resistance colder core of the element after the outer portion has been converted to liquid only partially.

It is not believed that this effect is caused by a variation in current density, as current density itself is a function of the total element cross section, although we do note a similar trend in average current density accompanying the phenomenon. A plausible explanation of this phenomenon lies in the following analysis.

If *n* similar conductors in parallel divide a given current equally and all fuse simultaneously, the total i^2t is given by the relation

$$\Sigma i^2 t = F_m d^4 n^2$$

where F_m is a constant (0.021 for silver) and *d* the wire diameter in mils. It is a significant fact, however, that should this current be allowed to fuse one conductor at a time in a continuous process until all *n* conductors are blown, the above relation then would be replaced with

$$\Sigma i^2 t = F_m d^4 n$$

Here the *n* is not squared. Now consider the *n* similar conductors in parallel within the fuse to be divided into two random parts, *r* and *q*, such that $(r + q) = n$. Now should *r* and *q* be fused in sequence, it follows that

$$\Sigma i^2 t = F_m d^4 (r^2 + q^2)$$

or

$$\Sigma i^2 t = F_m d^4 (n^2 - 2rq)$$

We are interested now in the per unit reduction in F_m as expressed by the ratio

$$F_m \left(\frac{n^2 - 2rq}{n^2} \right)$$

It is this fraction that we describe as an apparent reduction in F_m. For example, when *n* is 48 and we allow *r* to be progressively 0, 1, 2, 3, 4, ... and so on, the preceding fraction becomes 1.0, 0.96, 0.92,

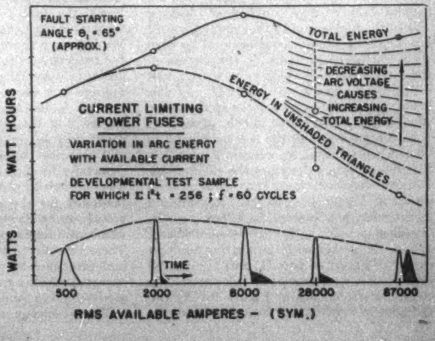

Figure 3. Reproductions of oscillograph watt element traces taken during interruption tests of five fuse samples of the same design at various available currents

Values of the time integrals (energy) of these traces are plotted at top

0.883, 0.848, ... and so on. Accordingly, the new F_m would take on apparent values of 0.0210, 0.0201, 0.0193, 0.0185, 0.0178, ... and so on.

This process within a fuse can be viewed as two separate phenomena, both of which we believe play a part in the observed reduction of F_m. The first may be that assumed previously, namely, the nonsimultaneous melting of the fuse wires. This is evidenced by the fact that F_m is observed to decrease as n increases. The second is skin effect in each and every conductor. The preceding numerical results would be duplicated if we considered each wire to be divided into 48 rings of equal area. If the outside 3 rings were assumed to fuse first, followed by the fusing of the inner 45 rings, then F_m would become the 0.0185 mentioned before. The actual skin effect phenomenon is, of course, not as discontinuous as suggested here, but a smooth variation which is only approximated in this analysis. That skin effect may play a part in the reduction of F_m is evidenced by the fact that F_m decreases as the wire diameter increases, wire diameter also increasing as n increases in practical fuse designs.

We agree with Gibson's conclusions concerning the expected performance of any powder-filled fuse which attempted to operate on the steep left-hand side of the energy characteristics of Figure 5, part II. As indicated in Figure 4 preceding, there is little current-limiting action in this region, and current zeros occur when the initial recovery voltage e_3 is near the peak of the generated voltage. In general, interrupters which operate in this region must be equipped to insulate the arc gap efficiently in order to cope with these peak recovery voltages and their attendant rates of rise. The short arc gap of a powder-filled fuse, which would be necessary to develop the low arc voltage, would be unable, we believe, to prevent the continued re-establishment of current. Fuses which utilize this energy to generate gases of high dielectric strength (expulsion fuses) can and do operate in this region, as indicated in Figure 4, part II. Such fuses are characteristically sensitive to the rate of rise of the recovery voltage.

The heating curves of Figure 8, part II, as Gibson points out, do not extend low enough to be useful for fuses of low current ratings, or for moderate current ratings at high values of short-circuit current. Accordingly, the curves of Figure 1 are included in this closure. These curves are a plot of the approximate equation 22 of part II and are expressed in radians instead of degrees. The upper portions of the curves of Figure 1 are corrected by about one per cent to align with the lower portions of the curves of Figure 8 of part II, which in turn are a plot of the exact equation 5 of part II. In this manner much of the error of the approximate equation 22 is eliminated in the curves.

In compliance with the last paragraph of Gibson's discussion, we submit Figure 2 in this closure which describes the effect of arc voltage magnitude upon the characteristics shown in Figure 10, part II. Retaining a rectangular arc voltage and holding the closing angle fixed at 60 degrees, equations 3, 5, and 6, together with the heating function curves of Figure 8, part II, and Figure 1 of this closure, were employed to compute the characteristic curves of Figure 2. As the arc voltage is decreased from 1.2 times the generated voltage crest, the arc energy rises. The peak liberated energy for a rectangular arc voltage equal to the crest of the generated voltage ($e_B = 1.0$) is slightly more than double the peak energy for an arc voltage crest of 1.2 times the generated crest ($e_B = 1.2$), and occurs at a current about 10 times as high as the peak energy current for the higher arc voltage.

This trend makes it imperative for the fuse designer to observe closely the characteristic arc voltage of his fuse in order to determine more accurately the current and closing angle which will produce the peak arc energy. His problems are complicated further if the arc voltage magnitude changes appreciably as the fault current is increased. The limited information given in Figure 10, part II, and Figure 2 of this closure should be considered only as sample characteristics, and do not express all the characteristics required to formulate a complete specification for the testing of fuses.

As the rectangular arc voltage decreases below the crest of the generated voltage, the arc energy increases rapidly as indicated in Figure 2 of this closure. Moreover, for arc voltages in this region the arc energy increases continuously with the megavolt-amperes (except for a very few cases having low closing angles and arc voltages in the region $e_B = 0.96$ to 1.0). It follows that any fuse which progressively produces lower arc voltages as the rate of rise of fault current is increased eventually will experience an upturn in arc energy as the megavolt-amperes are increased further. This will lead to eventual failure and hence dictate the ultimate interrupting capacity of the fuse. Such a trend is indicated in Figure 3 of this closure, discussed in the following text. The performance of the fuse at extremely high currents also is influenced considerably by the rate of rise of the arc voltage, a slow rate of rise augmenting the increase in arc energy.

Applications in circuits of various frequencies have demonstrated easier performance in the higher frequency circuits (reference 11 of part I). We have no experience with this fuse in circuits of very high frequencies (above 400 cycles), so are unable to answer Gibson's question on this point. We recommend that adequate tests be made to confirm any analytical prediction in this high frequency region.

Gibson suggests that a high rate of rise of fault current may induce breaking of the fuse tube. In our experience, the rate of rise of fault current affects only the melting time and the peak current, any tendency to break the fuse tube arising from energy associated with the arc current and voltage. We believe a more plausible explanation of the tendency for tubes to break at high fault currents is an inherent lowering of the arc voltage in a given fuse design as the fault current is increased. An examination of Figure 2 of this closure will reveal how a small decrease in the effective arc voltage will increase the arc energy considerably, especially for heavy currents. This effect was pointed out previously together with the tendency for the rate of rise of each arc voltage to be slower for high fault currents, which also tends to increase the arc energy. These two effects, both of which accompany high fault currents and give rise to increased energy, provide a rational explanation which merits further empirical study.

The arc energy can be recorded readily by use of an oscillograph watt element. Such records taken of five similar fuse samples tested at various available currents are shown in Figure 3 of this closure. The energy represented by the unshaded areas decreases with increasing available current beyond a maximum at 2,000 amperes, as would be expected from the trend of the curves of Figure 10, part II, which are on the basis that arc voltage is constant and essentially flat-topped. That it diminishes is indicated by the secondary energy areas (shaded). The resulting total energy thus increases at very high available currents as suggested in Figure 2 of this closure. The graphical study of this phenomenon (Figures 3 and 7 of part II) reveals the mechanism by which the double-humped energy curves are created as the arc voltage falls below the generated crest.

Factors Affecting Insulation Resistance of Large D-C Machines

Discussion and authors' closure of paper 46-157 by J. S. Johnson and Conrad Weil, presented at the AIEE summer convention, Detroit, Mich., June 24–28, 1946, and published in AIEE TRANSACTIONS, 1946, November section, pages 705–10.

R. W. Wieseman (General Electric Company, Schenectady, N. Y.): In the paper which Johnson and Weil presented at the summer convention Figure 1 showed a comparison of the authors' linear approximation of the slot area of d-c machines

$$1{,}500\left(\frac{\text{horsepower}}{\text{revolutions per minute}}+2\right)$$

and the slot area formula

$$\alpha\,\frac{\text{kilowatts}^{1/2}}{\text{revolutions per minute}^{1/2}}$$

given in reference 1 of this discussion. In this figure the authors' value of the approximate slot area varied as much as one-eighth to two times the actual value, whereas mine varied as much as one-eighth to eight times the actual value. In my discussion of this paper at Detroit it was pointed out that in our study of this problem we have not found a single d-c machine whose slot area computed from

$$\alpha\,\frac{\text{kilowatts}^{1/2}}{\text{revolutions per minute}^{1/2}}$$

deviated so much from the actual area as the original Figure 1 of the paper indicated. Instead of one-eighth to eight times the actual value it was only about one-half to $1^1/_2$.

Figure 1 of the paper has been revised so that the authors' approximate value of the slot area varies as much as 0.2 to 3.8 times the actual value, whereas mine varies as much as 0.2 to 4.5 times the actual value.

The ratio of the computed to the actual slot area of a large number of d-c machines (up to 6,000 kw) arranged on a linear scale

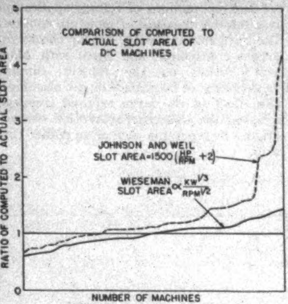

Figure 1

in increasing order is shown in Figure 1 of this discussion. The full line curve

$$\alpha\frac{\text{kilowatts}^{1/3}}{\text{revolutions per minute}^{1/2}}$$

varies from 0.6 to 1.45 times the actual value. The dotted curve

$$1{,}500\left(\frac{\text{horsepower}}{\text{revolutions per minute}}+2\right)$$

varies from 0.6 to 4.24 times the actual value, and this variation is nearly the same as that in the authors' Figure 1.

With a ratio of the computed to the actual slot area as much as 0.2 to 3.8, the authors' simplified formula will indicate an insulation resistance of some machines of 5 to 0.26 of the actual resistance. The authors apparently recognized this by advocating a minimum insulation resistance of two megohms when their calculated value falls below two megohms. A single minimum value of two megohms for all large machines is open to question. For a new machine I think it is too low.

In deriving an approximate formula for the insulation resistance of d-c machines, Johnson and Weil assumed that a number of quantities and ratios are constant. These assumptions and their variations are listed in Table I.

Although the authors' approximate formula based on the forementioned assumptions contains coefficients which were obtained from actual machines, it nevertheless will vary from five to one-fourth of the actual value. Thus I conclude that an expression for the insulation resistance of d-c machines based on the authors' simplified formula for slot area will not reflect the true proportions of some machines.

REFERENCE

1. INSULATION RESISTANCE OF ARMATURE WINDINGS, Robert W. Wieseman. AIEE TRANSACTIONS, volume 53, 1934, June section, pages 1010-21.

H. C. Marcroft (Pennsylvania Water and Power Company, Baltimore, Md.): This paper by Johnson and Weil is a timely contribution to the subject of electric machine resistance measurement, adding new facts concerning d-c machine insulation resistance to the information and test procedures already being used effectively for testing insulation of a-c machines.

The work done on a-c machines in trying to bring some order out of the mass of seemingly unrelated resistance readings demonstrated the need for consideration of absorbed moisture in the windings. To study the other variables the humidity problem was minimized by the practice of taking resistance measurements only on warm machines when they were shut down following a load period, because it was found that there was little meaning to resistance measurements taken on cold machines which had been absorbing moisture. This is of course of no value to the motor manufacturer.

The authors have seized a fine opportunity for evaluating humidity effects by their use of the many identical new machines coming from the production line. Humidity as well as temperature is a test variable which the authors have done well in emphasizing again, and it is hoped that the correction factors which have been applied so well to new machines may be adapted to machines in service also.

Moisture in insulation may be detected also by the dielectric absorption method using the ratio of resistance at 10-minutes time to the resistance at 1-minute time as the relative measure of moisture absorbed. This procedure does not establish any minimum value of resistance but evaluates the insulation in terms of polarization index. Values of polarization index have been established for large a-c machines, but it remains for this to be done for d-c machines. Experience shows that these values will be lower for d-c machines because of the greater number of creepage paths and less volume of insulation in these machines.

The linear formula for minimum resistance before proof voltage tests, which Johnson and Weil have suggested, is a definite contribution to the electric machine industry.

Caspar Goffman (Westinghouse Electric Corporation, East Pittsburgh, Pa.): I would like to comment, from a statistical point of view, on the validity of the resistivity corrections made in this paper.

The correction formulas used by the authors may be compared with formulas obtained by the method of least squares. I have made such comparisons, and in each case the corrections made by the authors compare favorably with corrections made using the more technical method.

For example, consider the reduction of resistivity to vapor pressures of 0.1 inch of mercury at the time of test and two months prior to test. The equation used in the paper is given by

$$\rho=cp_1^{-1}p_2^{-1} \tag{1}$$

where

ρ is the resistivity
p_1 and p_2 are vapor pressures at two months prior to test and at the time of test, respectively
c is a constant

The best linear fit and the best power law fits to the data are, respectively,

$$\rho=-5.2p_1-4.4p_2+4.67 \tag{2}$$

$$\rho=0.104p_1^{-0.748}p_2^{-1.07} \tag{3}$$

Equation 1 is a much closer fit to the data than equation 2 and is very nearly as good a fit as equation 3, the difference between equation 1 and equation 3 being well within usual limits of chance variation. In view of the fact that equation 1 is simple and that its difference with equation 3 may be attributed to change, it seems to be preferable to equation 3.

The justification of using any corrections at all lies in the fact that the standard deviation of corrected resistivities is about one-fourth of the standard deviation of corrected resistivities.

J. S. Johnson and **Conrad Weil:** The authors wish to point out a correction to Figure 1 of the paper as originally submitted. The curve on this figure showing the comparison of the correlation of the Wieseman size correction originally showed a one-eighth to eight extreme variation for the central 99 per cent of the machines checked. This value should be one-third to three.

Based on this corrected data as now shown in Figure 1, Wieseman's size correction shows approximately the same variability as the size correction derived by the authors. It should be pointed out that the authors' linear approximation compares favorably with both Wieseman's and the authors' theoretical corrections. The linear correction definitely has the advantage of simplicity.

The discrepancy between the authors' linear correction and the Wieseman size correction as shown in the curve of Wieseman's discussion could be explained by the choice of ratings used in checking the formula. Wieseman does not state how many machines were involved in his comparison nor the range in speed and horsepower of the designs checked. The authors' comparison included 200 representative

Table I

	Assumption		Variation
(a).	$\frac{\text{Horsepower}}{\text{Revolutions per minute}}=K_1D^2L$		K_1 varies about 2 to 1
(b).	Insulation thickness	(Constant)	Varies about $1^3/_4$ to 1
(c).	$\frac{\text{Slot area}}{\text{Creepage area}}$	(Constant)	Varies about 2 to 1
(d).	Slot depth	(Constant)	Varies about $1^3/_4$ to 1
(e).	Slot pitch	(Constant)	Varies about $1^3/_4$ to 1
(f).	$\frac{\text{Armature length}}{\text{Armature diameter}}$	(Constant)	Varies about $2^1/_2$ to 1

designs in the range of 100 to 7,000 horsepower. In the 100- to 1,000-horsepower range, low, medium, and high speed machines were checked for each standard frame size. In the 1,000- to 7,000-horsepower range, machines on which actual insulation resistance data were available were used for checking machine size corrections. The largest machines checked were 7,000-horsepower 35-rpm designs.

The paper points out that physical size of the machine is only one of a number of factors which affect machine insulation resistance. Our experience has been that ambient humidity conditions exercise an important influence on commutator type machines. The discussion of Marcroft also makes this point.

Marcroft mentions the use of the ratio of insulation resistance as measured after ten and one minutes of voltage application as a relative measure of moisture content. The authors have made similar checks on large a-c machines and agree with Marcroft's views on the subject. Similar checks have not been made on d-c machines. The effect of the large exposed creepage areas on the dielectric absorption characteristics in general is such as to reduce this ratio.

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